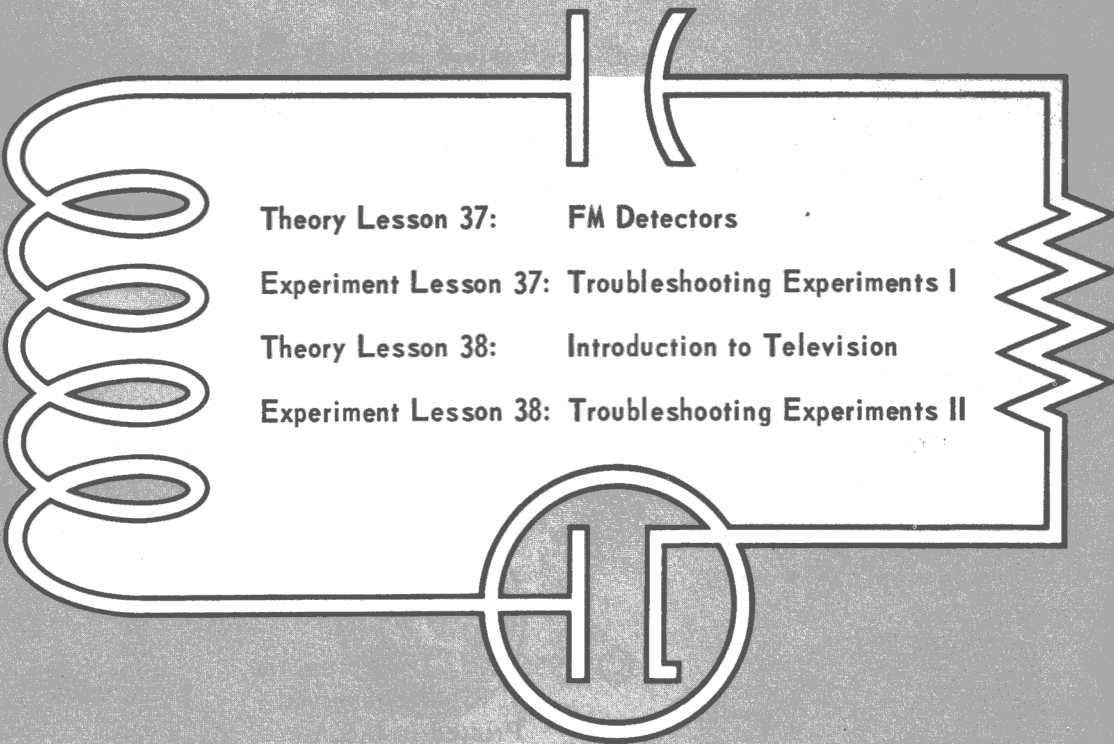


ELECTRONIC FUNDAMENTALS



Theory Lesson 37: FM Detectors

Experiment Lesson 37: Troubleshooting Experiments I

Theory Lesson 38: Introduction to Television

Experiment Lesson 38: Troubleshooting Experiments II

RCA INSTITUTES, INC.

A SERVICE OF RADIO CORPORATION OF AMERICA

New York,

N. Y.



ELECTRONIC FUNDAMENTALS

THEORY LESSON 37

FM DETECTORS

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Theory Lesson 37

INTRODUCTION

This lesson discusses FM receiver circuits, including the theory of operation of FM detectors, limiters, and i-f amplifiers. It provides the background you need in order to rapidly service troubles that may develop in these stages. When you get the Service Practices booklet you will learn about front end and automatic frequency control circuits used in FM sets.

A block diagram of an FM receiver is illustrated in Fig. 37-1. The incoming FM signal enters at the antenna input, is amplified by the r-f amplifier, and changed into an intermediate-frequency signal by the heterodyning of the oscillator with the r-f signal in the mixer section. The output of this section is an FM signal, but at the intermediate frequency. The i-f amplifier amplifies the incoming i-f signals; the limiter removes any amplitude modulation that may be present; and the FM detector extracts the modulation, or audio intelligence, from the FM modulated i-f signal applied to it. The audio signal output of the FM detector is applied to the audio stages, which amplify

it and feed it to the speaker. The power supply supplies the voltages needed to operate the various receiver stages.

37-1. SIDE TUNED DETECTOR

The FM detector is the most complex stage in an FM receiver. You have already learned something about FM detection from your study of slope detectors. Now you will learn more about FM detection, and study other types of FM detector.

Side-Tuned Detectors. Before we discuss any currently-used commercial FM detectors, let's consider the side-tuned detector; although it is no longer in common use, it is fairly easy to understand. Studying it will give you a basis for analyzing the operation of more complex FM detectors.

A *side-tuned detector* is shown in Fig. 37-2. Signal from the i-f amplifier plate circuit transformer winding (L_p) is coupled to two tuned secondary windings. The tuned secondary circuits connect to the inputs of two diodes functioning as detectors.

The detection action is similar to that occurring in a single slope detector. In this case, there are *two* slope detectors.

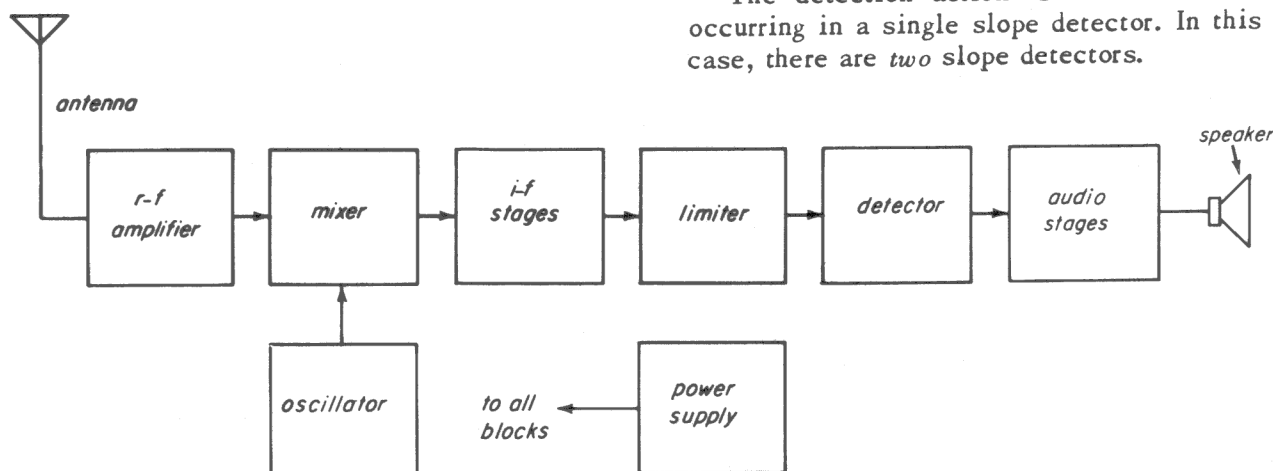


Fig. 37-1

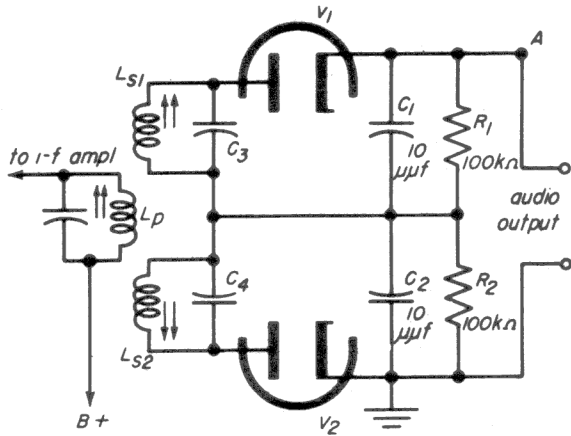


Fig. 37-2

Each tuned circuit converts the incoming FM i-f signals into amplitude-modulated i-f signals by its slope detection action.

Each secondary tuned circuit resonates at a different frequency. One resonates at a frequency *below* the FM i-f carrier (f_1), the other resonates at a frequency *above* it (f_2), Fig. 37-3). The resonant frequencies are separated an equal amount from the carrier frequency (f_c). Because of the difference in tuning, any i-f signal other than the carrier will produce a signal voltage in one tuned circuit different from the signal voltage it produces in the other.

This is illustrated in Fig. 37-4. At a frequency *below* the carrier, the amplitude of the signal voltage developed in the upper tuned circuit is a_1 ; the amplitude of the signal voltage developed in the lower tuned circuit is a_2 . At a frequency *above* the carrier, the amplitude of the signal voltage developed in the upper tuned circuit is c_1 ; the amplitude of the signal voltage developed in the lower tuned circuit is c_2 . When the unmodulated carrier is coming in at

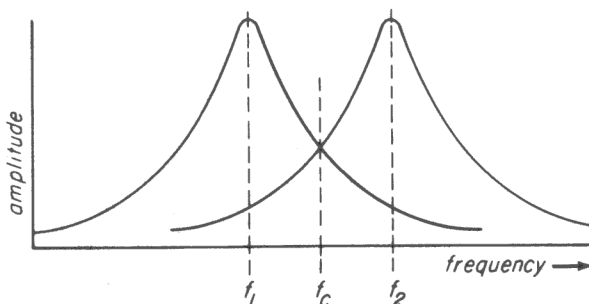


Fig. 37-3

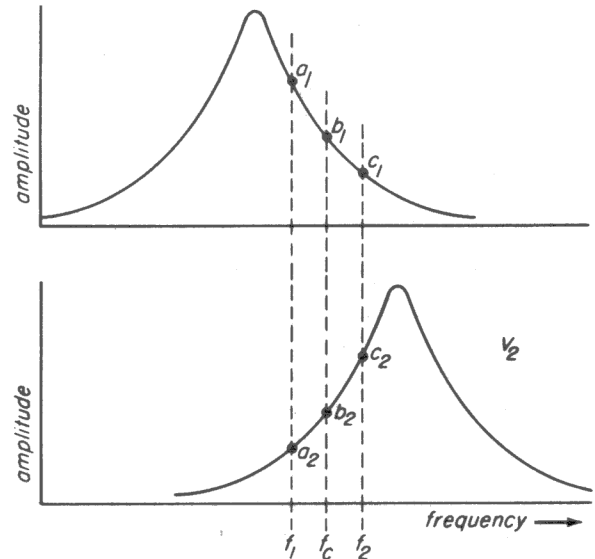


Fig. 37-4

the resting frequency (f_c), the signal voltage produced in both tuned circuits is the same ($b_1 = b_2$).

The diode detectors recover the audio intelligence from the amplitude-modulated i-f signals. Each diode conducts when the applied i-f signal makes its plate positive with respect to cathode. The detected output of the upper diode (V_1) appears across R_1 (see Fig. 37-5); the detected output of the lower diode (V_2) appears across R_2 . Capacitors C_1 and C_2 (Fig. 37-2) filter the i-f component out of the detector output.

simplified V_1, V_2 circuits

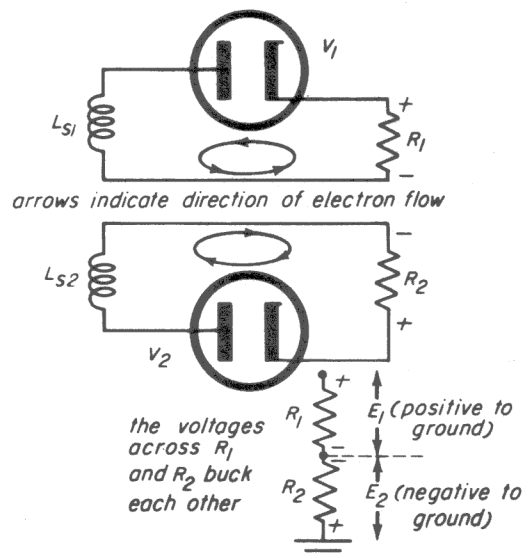


Fig. 37-5

The voltage developed across R_1 as a result of current flow through V_1 is *positive* with respect to ground; the voltage developed across R_2 as a result of current flow through V_2 is *negative* with respect to ground (Fig. 37-5). Therefore, the two voltages buck each other. The net voltage present at any instant between point A and ground is equal to the difference between the two voltages; the polarity of this net output voltage is the same as the polarity of the larger of the two voltages. For instance, if the larger voltage is +4 volts and the smaller one is -3 volts, the net voltage will be +1 volt. When the two voltages are equal, they cancel, and the output voltage is zero.

Since both diodes do not conduct at the same time, it might seem that the voltages across R_1 and R_2 neither appear simultaneously nor buck each other. But, in fact, the capacitors across R_1 and R_2 and the resistors themselves form relatively long time-constant networks. As a result of the discharge of the capacitor across it, the voltage across one resistor remains present long enough for the voltage across the other resistor to develop and buck it.

Figure 37-6 illustrates the output voltages produced when signals of several different frequencies are coming in.

Figure 37-6a shows the voltage outputs when the carrier is arriving. The input signal applied to one diode is equal to the input signal applied to the other diode at this time. The equal input signal voltages produce equal and opposite voltages across the diode load resistors. These voltages cancel. Therefore, there is no detector voltage output. Zero or no voltage output is desired at this time, since the FM carrier contains no audio intelligence, and should produce no detector voltage output.

When the incoming FM i-f signal drops below the frequency of the carrier, the amplitude of the signal voltage input to diode V_1 exceeds the amplitude of the signal voltage input to diode V_2 (Fig. 37-6b). Consequently, V_1 conducts more heavily than V_2 , causing the positive voltage output developed across R_1 to exceed the negative

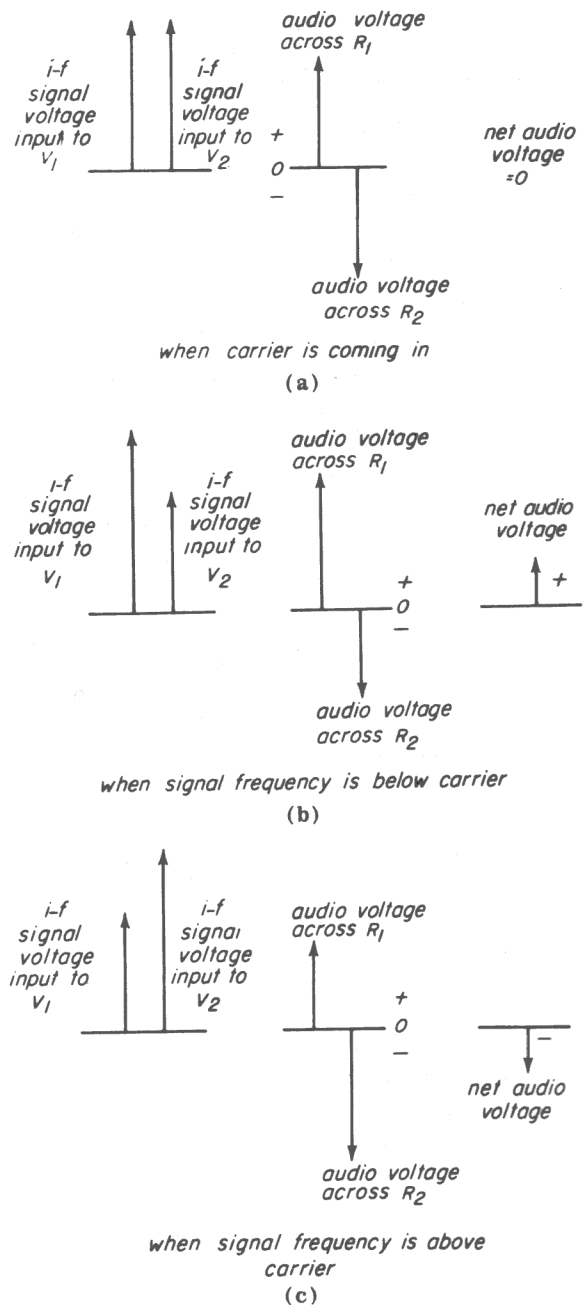


Fig. 37-6

voltage output across R_2 . The net audio output voltage is positive.

When the incoming FM i-f signal rises above the frequency of the carrier, the amplitude of the signal voltage input to V_2 exceeds the amplitude of the signal voltage input to V_1 . Therefore, V_2 conducts more heavily than V_1 , causing the negative voltage output developed across R_2 to exceed the positive voltage developed across R_1 . The net voltage output is negative.

The amplitude of the net voltage output at any instant is proportional to the frequency deviation of the incoming i-f signal from the carrier. The frequency deviation is thus converted into a proportional amplitude variation. The rate at which the amplitude variation is produced is the same as the rate of change of the FM i-f signal, which is, in turn, the same as the rate of change or frequency of the original audio signal. The frequency characteristic of the audio signal and its amplitude characteristic are thus

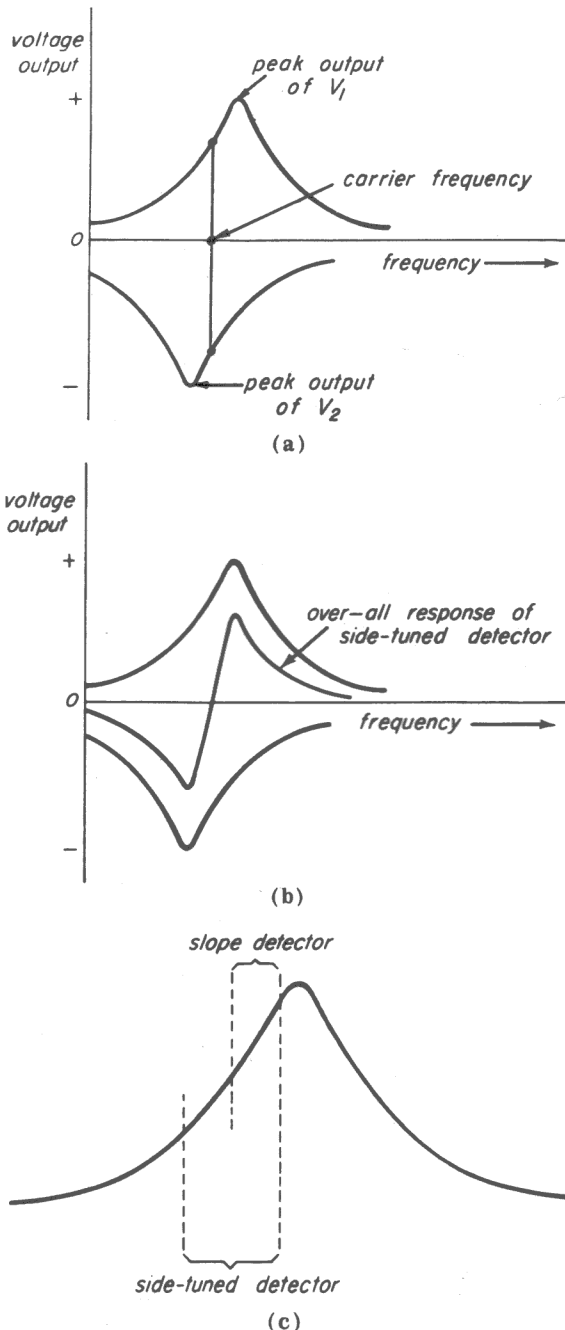


Fig. 37-7

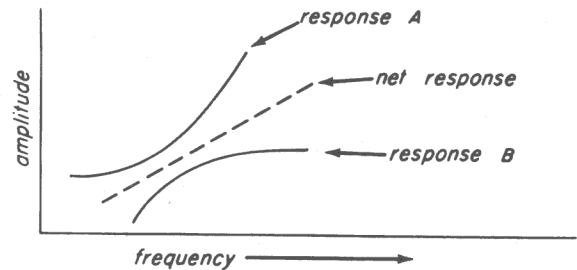


Fig. 37-8

reproduced at the detector output.

Advantages of Side-Tuned Detector Over Single Slope Detector. In order to understand the advantage of using two tuned circuits and a double diode detector, consider the graph in Fig. 37-7a. In this illustration the output of V_2 is at a peak below the carrier frequency; the output of V_1 is at a peak above the carrier frequency. Because signals applied to one tuned circuit produce detector output voltages opposite in polarity to those produced by signals applied to the other tuned circuit, the response of one diode section is inverted with respect to the other.

When the voltage output of the two diode sections combines, the over-all response of the circuit is as illustrated in Fig. 37-7b.

Note that half of the linear portion of this over-all response covers a wider range of frequencies than does the linear portion of the response of either diode section alone. How much longer the linear portion of the over-all response of a side-tuned detector is, compared to the linear portion of a single slope detector, is shown in Fig. 37-7c.

We can illustrate the way in which two nonlinear curves add together (as in Fig. 37-7b) to produce a linear curve by considering two simple response curves with opposite curvatures (Fig. 37-8). When these curves are added, a straight line or linear response results, since the curvature of one response is cancelled by the equal and opposite curvature of the other.

Loud sounds are represented by a wide frequency swing of the signal at the input to

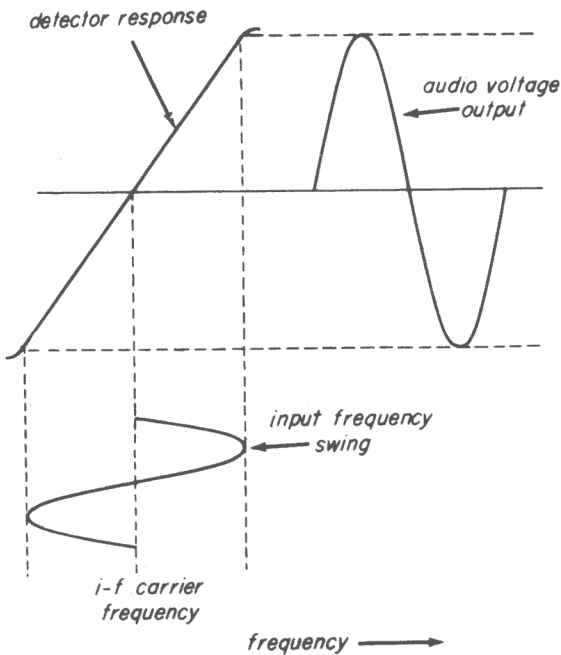


Fig. 37-9

the detector. As shown in Fig. 39-9 a wide linear response makes possible the detection of loud sounds. This feature is particularly important, since wide-band (wide frequency swing or deviation) FM is used by FM broadcasting stations.

Better rejection of amplitude modulation is provided by the side-tuned detector than by a single-tuned slope detector. When the carrier frequency is coming in, equal signal voltages are applied to both diodes, producing equal and opposite diode output voltages. Amplitude modulation associated with the carrier is therefore effectively canceled out. When signal frequencies other than the carrier are coming in, complete AM rejection does not take place in the detector.

Side-tuned detectors are not frequently used in modern receivers, because it is difficult to align their tuned circuits. A slight misalignment will seriously affect the linearity of the detector's response; the non-linearity of one tuned circuit's response will not exactly cancel that of the other, under such conditions.

37-2. REVIEW OF VECTORS

Adding the amplitudes of out-of-phase sine-wave voltages is more difficult than

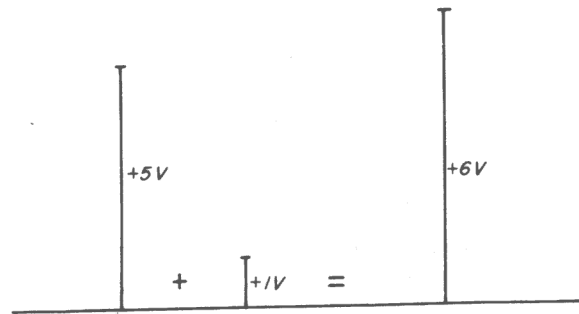


Fig. 37-10

adding either the amplitudes of d-c voltages or in-phase sine-wave voltages of the same frequency. Since a d-c voltage has an unvarying amplitude, all you have to do in order to add two d-c voltages is to add their amplitudes arithmetically. If one d-c voltage is +5 volts, and the other is +1 volt, and the two voltages are series aiding, the sum of the two is +6 volts (Fig. 37-10).

The addition of in-phase sine-wave a-c voltages of the same frequency is as simple as the addition of d-c voltages (Fig. 37-11a). If the peak amplitude of one voltage is 5 volts and the peak amplitude of the other voltage is 10 volts, the resultant (sum) of the two will be a sine-wave of the same frequency and phase, but having a peak of 15 volts. Sine waves A and B can be added instantaneously (as shown by points 1, 2, 3, 4, 5, 6 etc. in Fig. 37-11a), but the same result can be obtained more quickly by the

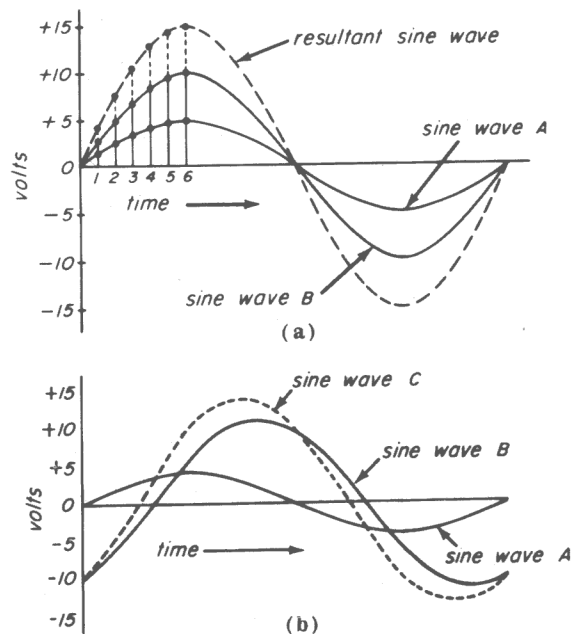


Fig. 37-11

direct addition of the two sine-waves. Peak values should be added – that is, the peak value of sine-wave *A* (5 volts) is added to the peak value of sine-wave *B* (10 volts) to obtain a third sine-wave (of peak value 15 volts). Effective (rms) values can also be used in adding sine wave voltages.

If you are trying to add two out-of-phase sine-wave voltages, the problem becomes more complex. In Fig. 37-11*b* sine wave *C* is the sum of out-of-phase sine waves *A* and *B*. As you can see in Fig. 37-11*b*, when two waveforms are out of phase, the simple system of addition cannot be used. At times in the cycle when both waveforms are positive-going, their amplitudes add. When one waveform is positive-going while the other is negative-going, the smaller of the two instantaneous amplitudes is subtracted from the larger one to determine the amplitude as well as the polarity of the resultant waveform at this instant. If the amplitude of one waveform is +5 volts, while the amplitude of the other is –1 volt, the amplitude of the resultant waveform at the instant under consideration is 4 volts; its polarity is positive. Thus, additions must be made at many points throughout a voltage or current cycle, making the graphical method of adding a-c voltages or currents time-consuming and tedious. The vector method of adding a-c voltages or currents is much simpler and faster.

You have already learned something about vectors. You know that vectors are lines that show both magnitude and direction. Therefore, vectors can be used to represent forces exerted in particular directions. The magnitude of the vector represents the magnitude of the force and the direction of the vector represents the direction of the force.

Figure 37-12*a* shows two vectors, both of which start from a common origin, point *O*. One vector represents a force of 75 pounds exerted in a northward direction. This vector has a certain length and points north. The other vector, which represents a force of 100 pounds exerted in a southward direction, is proportionally longer and points south.

When the smaller vector (shorter line) is subtracted from the larger vector (longer

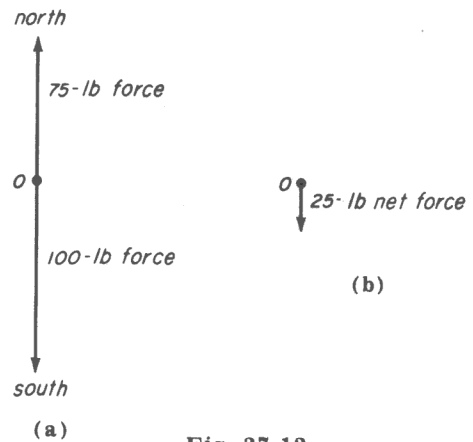


Fig. 37-12

line), the resultant vector (Fig. 37-12*b*) has a length that represents a force of 25 lbs. Since the longer vector pointed south, the resultant vector points south.

Vectors do not always point in opposite directions. For instance, the angle between two forces might be 90 degrees; then the vectors would form an angle of 90 degrees. In Fig. 37-13*a* two vectors of equal magnitude are at an angle of 90 degrees with respect to each other. The two vectors can be considered to be two sides of a parallelogram. The parallelogram can be completed as shown by the dashed lines in Fig. 37-13*b*. If a diagonal line is drawn from the origin *O* to the opposite point of the parallelogram as in Fig. 37-13*c*, the diagonal line will represent the vector that results when vector *A* and vector *B* are added. The length of the diagonal line represents the magnitude of

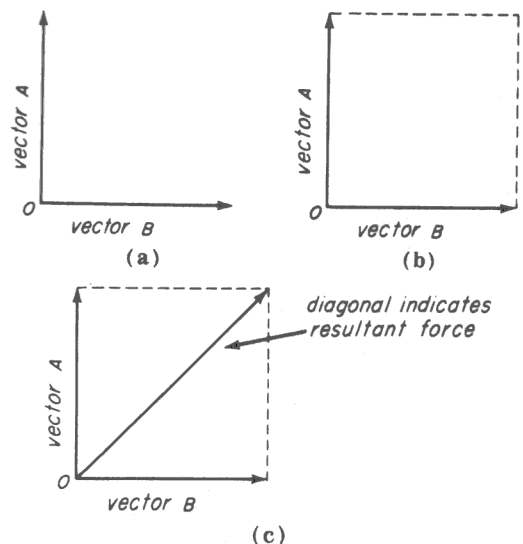


Fig. 37-13

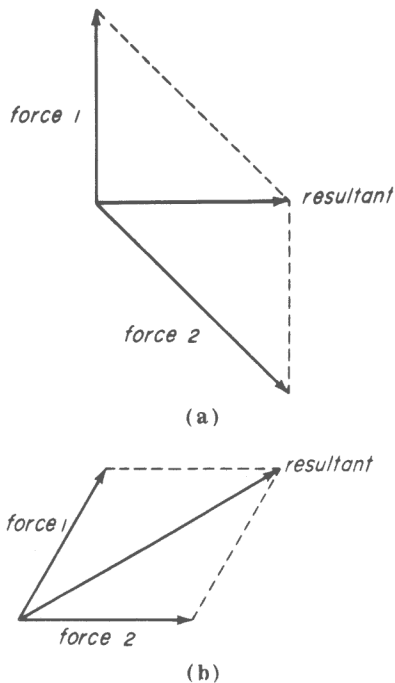


Fig. 37-14

the resultant vector; the direction of the diagonal is the direction of the resultant vector.

Thus, if vectors *A* and *B* are each one inch long, and vector *A* points north and vector *B* points east, the resultant vector will be 1.4 inches long, and will point in a northeast direction. Additions of other vectors are illustrated in Fig. 37-14.

Vectors are used to represent sine-wave voltages and currents. The length of a vector that is used to represent a sine-wave voltage or current indicates either the peak or the effective value of the voltage or current.

For reasons that will soon become plain, it is considered that a vector rotates. By convention, it is considered to rotate in a counterclockwise direction around its point of origin. When the vector has rotated 360 degrees, it has completed one cycle of rotation.

The vector is considered to rotate at the same frequency as the sine wave it represents. In other words, it completes one cycle of rotation in the same time that the sine wave completes one waveform cycle.

A vector with a length representing the peak value of a sine wave is shown in Fig. 37-15a. The vector is at the 0-degree point in its cycle. In the graph to the right of the vector, a sine wave is about to begin. The sine wave will follow the course shown by the dashed lines. In *b* of the figure, the vector is shown rotated 22.5 degrees from its starting position, and amplitude of the sine wave at the 22.5-degree position is shown. In *c* the vector and sine wave are at the 45-degree position, and in *d* both are at the 90-degree position. Note particularly that the distance of the vector above (or

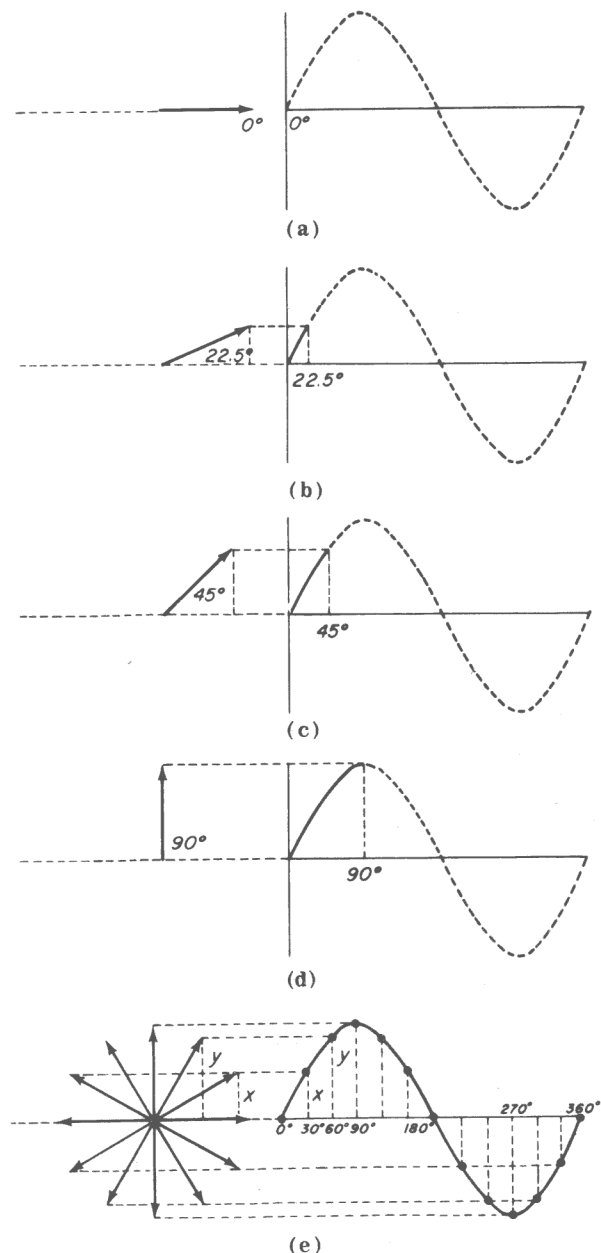


Fig. 37-15

below) the horizontal line at any phase angle in its cycle is equal to the amplitude of the sine wave at the same phase angle. In *e* is shown one complete rotation of the vector and one cycle of the sine wave. The vector makes one revolution while the sine wave goes through one cycle. Unless otherwise stated, a vector that rotates is assumed to rotate counterclockwise.

When a vector is used to represent a sine wave, it is not necessary to draw a separate vector for each point in the sine-wave cycle. Instead a vector is drawn at *one* point of the cycle, usually the peak value, of the sine wave. Since the vector does not change in length, and you know that it rotates as the sine wave moves, a single vector is enough to represent the entire sine wave.

When vectors are used to represent two out-of-phase sine waves of the same frequency, the vectors are drawn at the same phase angle as the sine waves. Suppose the two sine waves shown in Fig. 37-16a are to be shown vectorially. Vector *A* (Fig.

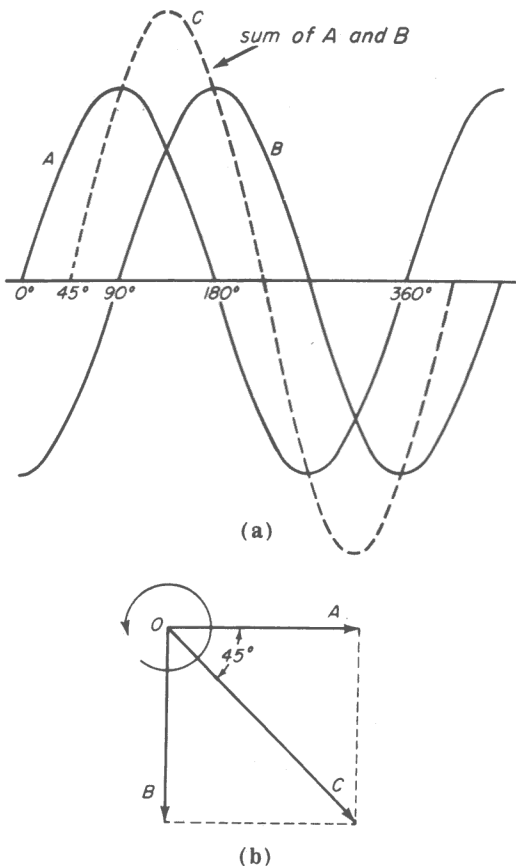


Fig. 37-16

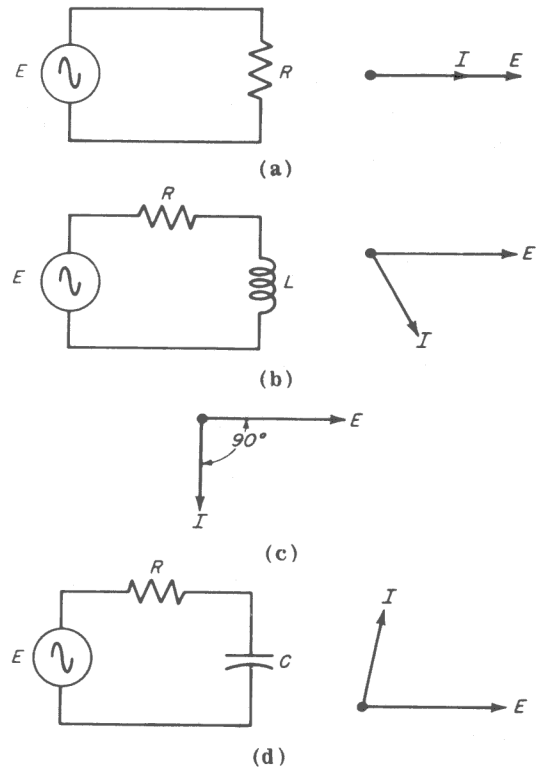


Fig. 37-17

37-16b) represents sine wave *A*. Vector *B* represents sine wave *B*. Since sine wave *B* lags sine wave *A* by 90 degrees, vector *B* lags vector *A* by 90 degrees. Since vectors are considered to rotate counterclockwise, vector *A*, which is ahead of vector *B*, must be represented as further *counterclockwise* than vector *B*.

Vectors *A* and *B* can be added as shown in Fig. 37-16b. The resultant, vector *C*, is equivalent to sine wave *C* in Fig. 37-16a. Note that vector *C* lags vector *A* by 45 degrees, and leads vector *B* by 45 degrees, just as sine wave *C* lags sine wave *A* and leads sine wave *B* by 45 degrees.

Other examples of the use of vectors to show phase difference are shown in Fig. 37-17. In this figure, the vectors are intended to show only phase relationship, not amplitude. Therefore, the vectors are not drawn to scale.

The phase relationship between current and voltage in a resistive circuit is shown by the vectors in Fig. 37-17a. Since *I* and *E* are in the same direction, the vectors that

represent them point in the same direction. In order to show both vectors pointing in the same direction, the two vectors must be drawn so that one is on top of the other. If both vectors were the same size, one would be completely covered by the other. In order to show both vectors, E has been made arbitrarily longer than I .

In a predominantly inductive circuit (Fig. 37-17b), current in the circuit lags the applied voltage as shown vectorially. The voltage vector starts its cycle of rotation earlier than the current vector. So, the voltage vector is further counterclockwise than the current vector. The angle between the vectors, called the *angle of lag*, depends on the ratio of inductive reactance to the resistance. The greater the amount of resistance, the smaller is the angle of lag. If the circuit were purely inductive and the resistance was zero, current would lag the applied voltage 90 degrees, as shown in Fig. 37-17c.

In a circuit which is predominantly capacitive (Fig. 37-17d), current *leads* the applied voltage, as shown vectorially in the figure. The angle of lead depends on the ratio of capacitive reactance to the resistance present in the circuit. The greater the amount of resistance present in the circuit, the smaller is the angle of lead. If the circuit were purely capacitive, and the resistance zero, the current would lead the applied voltage 90 degrees.

37.3. THE FOSTER-SEELEY PHASE-SHIFT DISCRIMINATOR

One of the two circuits commonly used for FM detection is the Foster-Seeley discriminator. A simplified circuit of such a discriminator is illustrated in Fig. 37-18a.

The limiter is coupled to the discriminator through the discriminator transformer T as well as a capacitor (C_3). The primary and secondary windings of the transformer are tuned to the same center frequency. The secondary of the transformer is center-tapped. The two halves of the secondary are labeled L_{S1} and L_{S2} in Fig. 37-18a. The

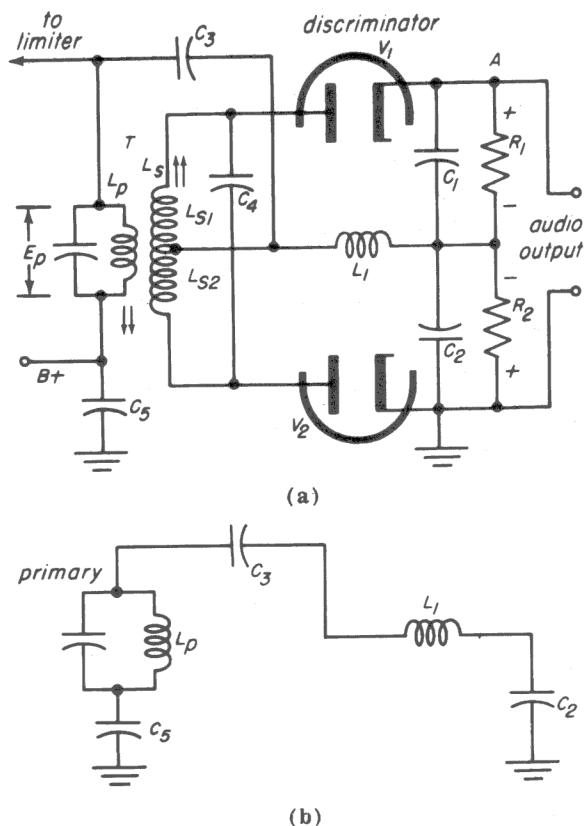


Fig. 37-18

entire secondary is designated L_S . The discriminator is generally a duo-diode tube. The diode sections (V_1 and V_2) have the same characteristics. The components in each diode circuit are also equal: R_1 equals R_2 , C_1 equals C_2 , and L_{S1} equals L_{S2} .

FM i-f signal is developed across the primary of the transformer in the limiter plate circuit and is applied between plate and cathode of each diode of the discriminator in two ways: capacitively, by way of C_3 , and inductively, via coupling from the primary to the secondary of the transformer.

Let's consider the capacitive coupling (Fig. 37-18b). The FM i-f signal voltage is across the primary of the discriminator transformer. The lower end of the transformer is grounded as far as the signal is concerned by the bypass capacitor C_5 . The upper end of the transformer is connected, through C_3 , to one (left) end of L_1 . The other (right) end of L_1 is grounded by C_2 . Thus, the right end of L_1 is connected effectively to the bottom end of the primary. Because the reactances of C_2 , C_3 and C_5 to

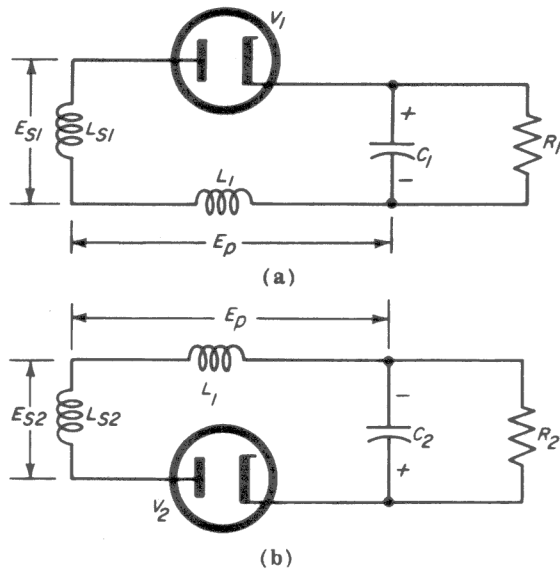


Fig. 37-19

i-f signals is so small, they may be considered short-circuits for these signals. Practically all of the signal voltage developed across the primary appears across L_1 .

Now that we have studied the capacitive coupling, let us analyze the inductive coupling. Since the primary L_P is inductively coupled to the secondary L_S , the FM i-f signal current in L_P will induce a voltage into L_S . Note that the secondary is center-tapped. As shown in Fig. 37-19a, as far as the upper diode V_1 is concerned, the upper half of the secondary voltage and the primary voltage across L_1 are applied to the diode. As shown in Fig. 37-19b, the lower diode V_2 has impressed across it the voltages across the lower half of the secondary and the primary voltage across L_1 .

The plate current of V_1 flows from cathode, to plate, through L_{S1} , through L_1 , through the load made up of resistor R_1 and capacitor C_1 in parallel, back to cathode.

The plate current of diode V_2 flows from cathode, to plate, through L_{S2} , through L_1 , through R_2 and C_2 in parallel, back to cathode.

The two diodes act, in combination, as a detector. The audio output of each diode appears across its load resistor and capacitor. The amount of voltage built up

across R_1 and C_1 and R_2 and C_2 depends on the voltage input to V_1 and V_2 , respectively. The net audio output voltage is made up of the sum of the voltages developed across R_1C_1 and R_2C_2 . These voltages are opposite in polarity (Fig. 37-19) and buck each other, just as in the case of the side-tuned circuit. The net voltage between point A and ground (Fig. 37-18a) and the polarity of this voltage depend on which voltage is larger.

Tuned Transformers. Tuned transformers play an important part in the operation of the discriminator. Figure 37-20a shows the discriminator transformer removed from the circuit. Let's discuss the details of the operation of this transformer. In order to simplify the explanation, we will disregard the fact that L_S is divided into L_{S1} and L_{S2} . We will refer to L_S only.

The transformer is tuned to the center i.f. — usually 10.7 mc. For the purposes of this discussion, let's assume that only the center i.f. (10.7 mc) is being applied to the circuit.

The voltage across the primary of the transformer, E_p , is shown vectorially in Fig. 37-20b. The current through the pri-

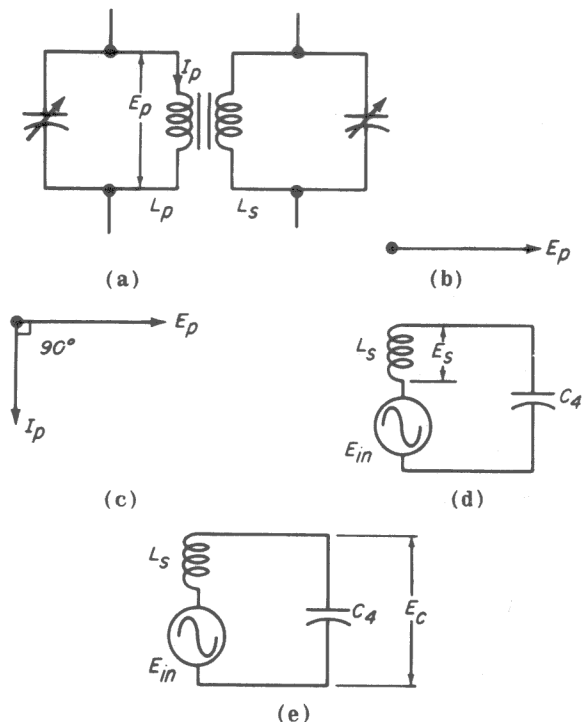


Fig. 37-20

mary, I_p lags E_p by 90 degrees (Fig. 37-20c). When I_p flows through the primary, a voltage E_{in} is induced in the secondary winding. The voltage induced in the secondary produces the same effects in the secondary circuit as would a generator in series with the secondary tuned circuit. Therefore, it is possible to represent the secondary circuit as shown in Fig. 37-20d and to call the voltage induced in the secondary E_{in} . This induced voltage, E_{in} , causes a current I_s to flow in the tuned secondary circuit.

Current I_s flows through the secondary coil, which has a reactance X_{LS} . The flow of the current through X_{LS} causes a voltage drop ($I_s X_L$) across L_s ; we will call this voltage E_s . The flow of I_s through X_C (the reactance of C_4) causes a voltage $I_s X_C$ to be built up; we'll call this voltage E_c (Fig. 37-20e).

Now we can use vectors to depict the entire process. Let's go back to the voltage across the primary. We can represent that voltage, E_p , by a vector (Fig. 37-21a). We can also represent the current I_p in the primary circuit by a vector that lags the voltage vector by 90 degrees.

Now let us use vectors to show what happens in the secondary. E_{in} , the voltage induced in the secondary, is either 180 degrees out of phase with E_p , or in phase with E_p , depending on the polarity in which the secondary winding is connected. (This is the case whether the primary and secondary circuits are tuned or untuned.) Assume that the winding is connected in such a way

that E_{in} is 180 degrees out of phase with E_p (Fig. 37-21a).

The secondary is at resonance. Therefore, the inductive and capacitive reactances in the secondary tuned circuit are equal and opposite. Since the reactances cancel out, the only opposition to the flow of current is the coil resistance. Therefore, the current I_s is phase with the induced voltage, E_{in} , (Fig. 37-21b).

When I_s flows through L_s , it causes a voltage drop across the coil. This voltage, E_s , leads I_s by 90 degrees (Fig. 37-21c).

Placing the E_s vector 90 degrees ahead of I_s causes E_s to be represented as in phase with I_p .

Up to this point, we have been considering L_s as a single secondary. But in fact, L_s is center-tapped. Center-tapping divides L_s into L_{s1} and L_{s2} . This splits E_s into two equal component voltages, E_{s1} and E_{s2} . E_{s1} is measured from the plate of diode V_1 to the center tap; E_{s2} is measured from the plate of diode V_2 to center tap. Since the voltage at one end of a transformer is opposite in polarity to the voltage at the other end, E_{s1} and E_{s2} are opposite in polarity. Therefore E_{s1} and E_{s2} are represented by vectors that are 180 degrees out of phase, as shown in Fig. 37-21d.

Recapitulating and referring to Fig. 37-19, the voltages applied to the upper diode V_1 are E_{s1} and E_p ; the voltages applied to the lower diode V_2 are E_{s2} and E_p . Since the two voltages that are applied to each diode are out of phase, it is the vector sum of E_{s1} and E_p that is applied to V_1 , the vector sum of E_{s2} and E_p is applied to V_2 .

In explaining the operation of the discriminator, it is necessary to consider its action at the carrier frequency, above the carrier frequency, and below the carrier frequency.

You have learned that when a single audio note modulates an FM carrier, the carrier deviates away from the resting fre-

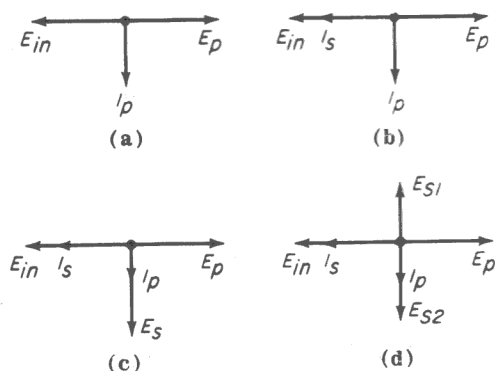


Fig. 37-21

quency. As the a-f signal goes through its positive half cycle, the carrier deviates in one direction, attaining the maximum deviation when the a-f signal reaches its peak value. When the a-f signal goes through the negative half cycle, the deviation away from the center frequency is in the other direction. Maximum (reverse) deviation occurs at the time the a-f cycle reaches the peak negative value.

Operation of the Discriminator at the Center Frequency. As we have explained, the i-f input voltage applied to the upper diode circuit is always the vector sum of E_{S1} and E_p . This is represented by E_{V1} in Fig. 37-22a. The i-f input to the lower diode circuit is the vector sum of E_{S2} and E_p , or E_{V2} . As you can see from the vector diagrams, the *magnitudes* of E_{V1} and E_{V2} are the same at the carrier frequency because this is the frequency at which the transformer is resonant. That means that the *same amount* of voltage is applied to the input of both diode circuits at the carrier frequency.

The voltages developed across R_1 and R_2 (Fig. 37-18) will be rectified, of equal amplitude, and opposite polarity under these conditions, making the total d-c voltage output zero. If C_1 and C_2 weren't in the circuit to filter the pulsations out of the rectified output voltage and there were no stray or interelectrode capacitances, the rectified i-f output of the discriminator would have the waveform shown in Fig. 37-22b. Note that the half-cycle waveforms are 90 degrees out of phase. It can be graphically shown that the addition of these waveforms produces an a-c waveform with a zero d-c voltage level. Therefore, the total d-c voltage output of the discriminator is zero, even though the a-c input voltages at the carrier frequency are 90 degrees out of phase.

With i-f filter capacitors C_1 and C_2 in the circuit, the pulsating d-c voltage at the output of each diode is smoothed out at a level equal to the peak value of the rectified i-f voltage pulses. (The importance of this fact will be made clear when we con-

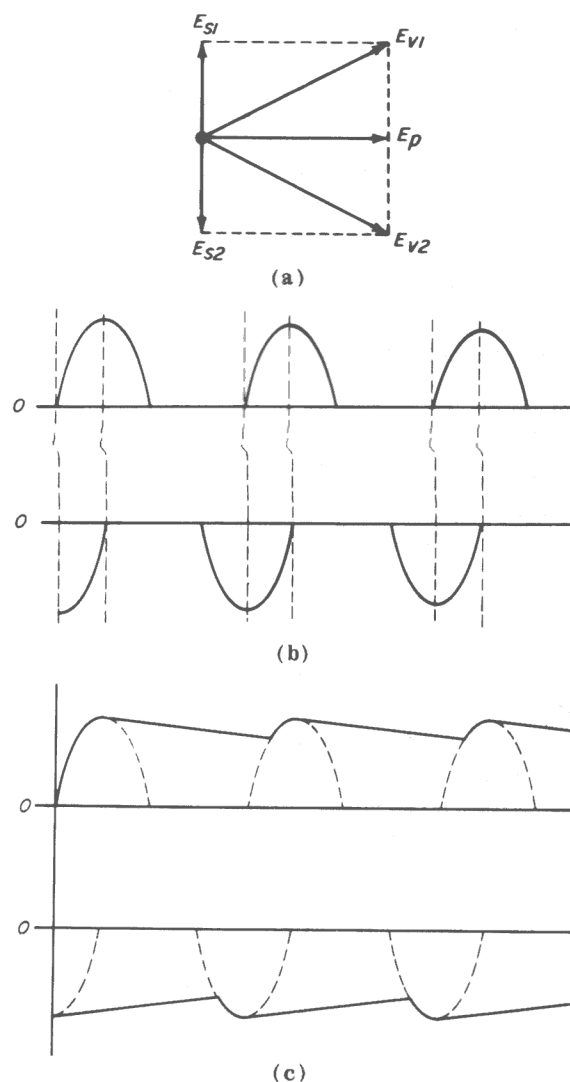


Fig. 37-22

sider operation above center frequency.) The filtered d-c voltages across R_1 and R_2 are equal and opposite, so the voltages still cancel, and the d-c voltage output is zero at the instant when the a-c input is at carrier frequency. This is a fundamental requirement of FM detection—that when the FM modulated carrier is at center frequency, the audio signal which is made up of a sequence of instantaneous d-c voltage levels of the rectified r-f signal should be at the zero-volt level in its cycle; therefore, the detector output should be zero volts d.c. at this time.

When the FM carrier is modulated (Fig. 37-23), the instantaneous d-c voltage across R_1 is *not* always the same as the corresponding instantaneous d-c voltage across R_2 . The difference in the instantaneous

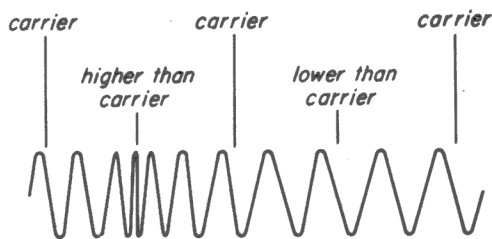


Fig. 37-23

d-c voltages across each of these resistors corresponds to the instantaneous amplitude of the audio signal.

Discriminator Circuit Action When Signal Frequency is Higher Than Center Frequency. When the incoming i-f signal is higher or lower in frequency than the one to which the secondary tuned circuit is resonant, the input voltages to each diode become unequal. Let's consider the vector relationship when the signal frequency is higher than the carrier.

At a frequency above the one to which the secondary tuned circuit resonates, the inductive reactance is greater than the capacitive reactance. The secondary circuit becomes inductive, since X_L is larger than X_C . I_S , the current flowing through the circuit, will lag the induced voltage, E_{in} .

The vectors are no longer in the same position as they were at the center frequency (Fig. 37-24a). Instead the I_S vector will move away from E_{in} (Fig. 37-24b). When the I_S vector moves, the E_{S1} and E_{S2} vectors will move in step with it (Fig. 37-24b). This is so because E_S (the total voltage across the secondary of the discriminator transformer) always leads I_S by 90 degrees (since the voltage across a coil leads the current in it by 90 degrees). E_{S1} and E_{S2} are simply the two halves of E_S .

E_{V1} , the sum of vectors E_{S1} and E_P , and E_{V2} , the vector sum of E_{S2} and E_P , are shown for a frequency above center in Fig. 37-24b. Note that E_{V1} is larger than E_{V2} . This instantaneous condition exists at the time the input waveform is going through the portion of the cycle labeled *higher than carrier* in Fig. 37-23. Since the input to V_1 is greater than the input to V_2 , V_1 conducts more heavily than V_2 . There-

fore, the rectified d-c output voltage of V_1 will be greater than the rectified d-c output voltage of V_2 . Therefore, the peak-level filtered d-c voltage output of V_1 will be higher than that of V_2 (Fig. 37-24c). The net voltage at the discriminator output, which is the difference between the filtered d-c voltage across R_1 and R_2 , will be positive in this case. Thus, an FM signal that deviates above the carrier frequency causes instantaneous positive output voltages to be developed. The greater the deviation (in this case an increase in frequency), the greater will be the instantaneous positive voltage. The d-c output will be proportional to the full peak value of the rectified pulses because capacitors C_1 and C_2 filter at the peak level. Thus, although the circuit can work without C_1 and C_2 , these capacitors ensure that maximum possible output will be delivered and that there will be no r-f variations in the d-c output.

Operation of the Discriminator Below the Center Frequency. When the incoming i-f signal falls below the frequency to which the secondary circuit is tuned, the capacitive reactance in the tuned circuit is greater than the inductive reactance. Thus, the secondary is capacitive. Current leads the applied voltage — that is, I_S leads E_{in} (Fig. 37-24d). The I_S vector moves below E_{in} and E_{S1} and E_{S2} rotate with it.

E_{V1} is the sum of vectors E_{S1} and E_P ; E_{V2} is the sum of vectors E_{S2} and E_P . E_{V2} is greater than E_{V1} and will cause the lower diode V_2 to conduct more heavily than the upper diode, V_1 . In turn, there is a greater rectified voltage output from V_2 than from V_1 (Fig. 37-24e). Therefore there is a greater filtered d-c voltage drop across R_2 , and a lower d-c voltage drop across R_1 . The net voltage output of the discriminator will be negative in this case. Hence, an FM signal deviation lower than center frequency develops an instantaneous output voltage that is proportional to the deviation.

Cancellation in the Discriminator Output. So far we have been discussing cancellation without considering what the process in-

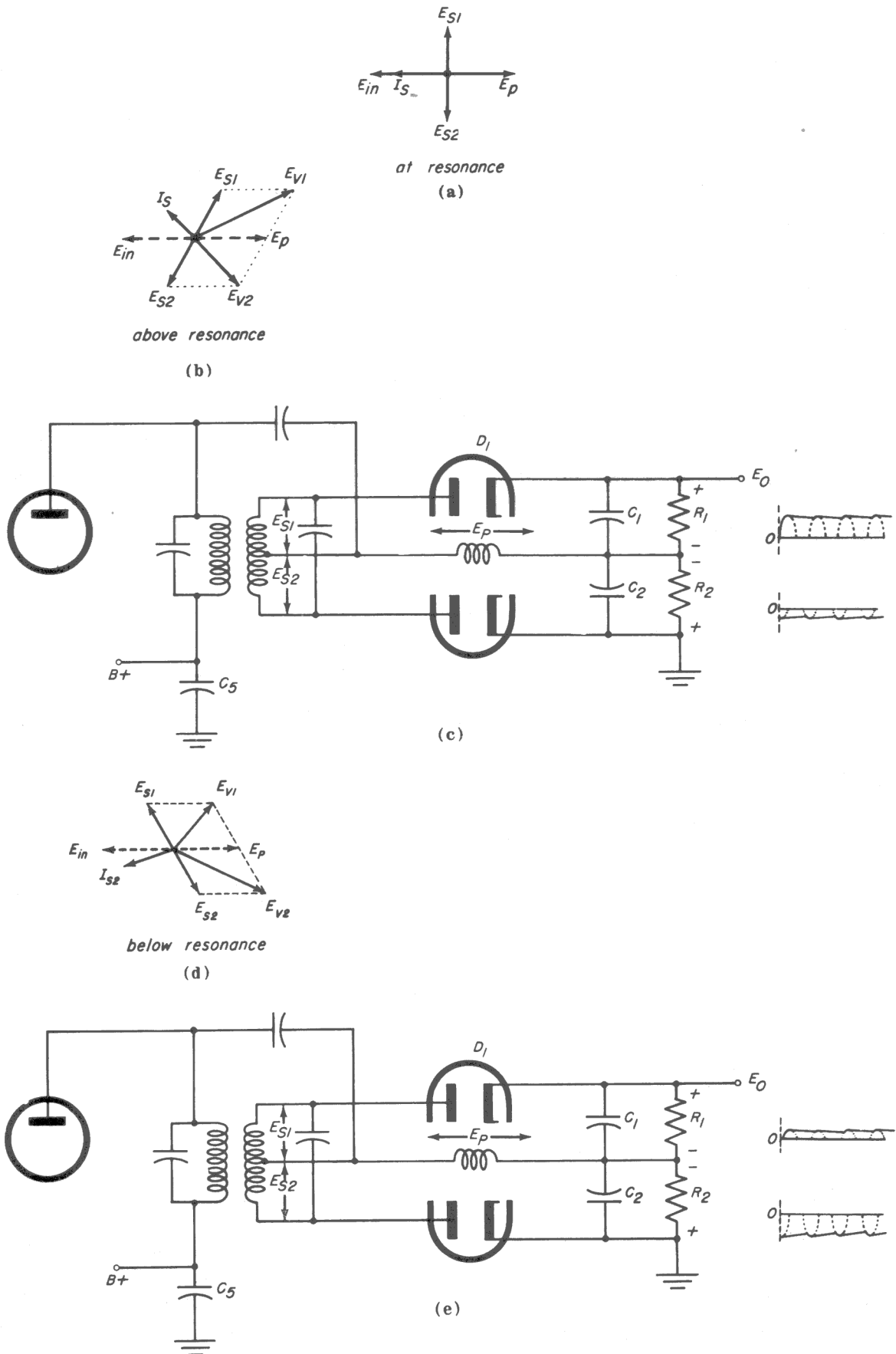


Fig. 37-24

volves. But unless we explain the process in detail, you will find the numerical results of cancellation confusing. You must understand the cancellation in order to know how the audio output signal is formed.

We have already pointed out that the *instantaneous* output of the discriminator is the d-c voltage that results when the peak d-c voltage across one load resistor is subtracted from the peak d-c voltage across the other load resistor. At the carrier frequency, when there is no modulation of the FM carrier, the d-c voltage across one load resistor, is equal to the d-c voltage across the other load resistor as shown in Fig. 37-25a, and complete cancellation occurs.

Figure 37-25a shows the output of the discriminator for an extended period of no modulation. During this period, the d-c voltages, being equal and opposite, cancel out. In actual operation of an FM receiver, however, the period during which there is no modulation may be much shorter than the interval shown in Fig. 37-25a. There may be no modulation only during brief intervals between modulated periods, as shown by the points marked X in Fig. 37-25b. During the intervals in which the FM carrier is modulated, the d-c output across the load resistors varies from instant to instant. When the i-f input to the discriminator is above the carrier frequency, the instantaneous d-c output level rises from what it is when the i-f input is at the carrier frequency. When the i-f input is below the carrier frequency, the instantaneous d-c output level falls below the output level at the carrier frequency. Figure 37-25b shows the output across the two load resistors of a discriminator when the modulation on the FM carrier is a sine-wave cycle.

It is when the carrier is modulated that the problem of how to calculate the effects of cancellation arises. In Fig. 37-25c, the waveforms of part b of the figure are reproduced. With them is shown a simple circuit that could be used to produce these waveforms. By considering this circuit, you will gain insights concerning the operation of the discriminator. Notice that the circuit consists of two separate voltage sources. Each of these separate sources consists of

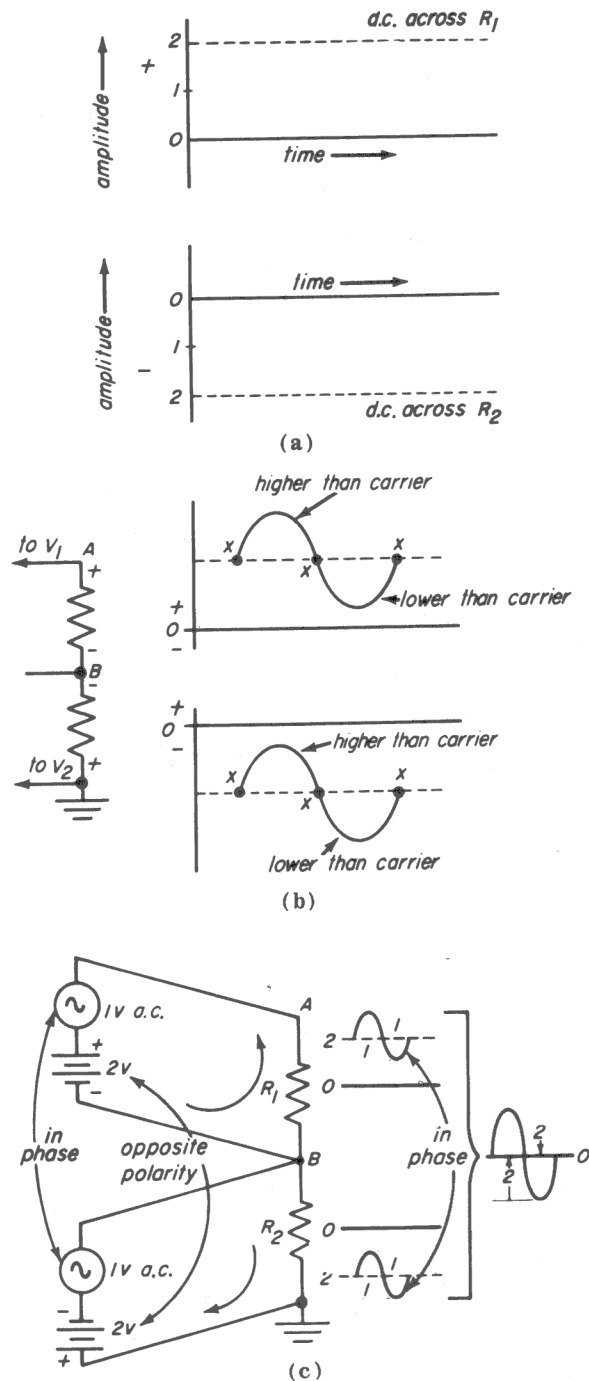


Fig. 37-25

a battery to produce d.c. and a generator to produce a.c. As a result, each of the sources provides an a-c voltage varying above and below a d-c reference level. Such an output is similar to the output of the discriminator circuit, as shown in b of the figure, in which the instantaneous output voltages of the discriminator constitute an a-c voltage varying above and below a d-c reference level.

Each of the voltage sources is connected across its own load, and the loads of each of the two sources are connected in series. This too is the case in the discriminator circuit you have been studying. The ground point is in the same relative position in this simple circuit as it is in the discriminator. There are differences between the discriminator circuit and the simple circuit we are discussing too. But the important point about these circuits so far as this discussion is concerned is that both types of circuit can be used to produce a similar output. The output across the load resistors of both circuits is a combination of d-c voltages of opposite polarity and a-c in-phase voltages. The two a-c sine waves are in phase even though current is flowing through the load resistors in opposite directions. Both waves go positive at the same time. In the case of the lower wave, going positive means going less negative.

Let's calculate the total voltage across the two load resistors (between point A and ground) when the two sine waves are at their most positive. The voltage across R_1 alone, as shown by the waveform net to R_1 , consists of a d-c reference level of +2 volts and an a-c component. The a-c component at this instant is +1 volt. Therefore, the total instantaneous voltage across R_1 is +3 volts. Meanwhile, during the instant that the upper sine wave has risen to its positive maximum of +3 volts, the current through R_2 has fallen to a minimum. As a result, the waveform for R_2 is at its lowest negative level. As you can see in the figure, this means that the lower waveform is at -1 volt. When you combine the two voltages, +3 - 1 volt = +2 volts. Thus, as the resultant waveform on the right hand side of the figure shows, the maximum positive output between point A and ground is +2 volts. You can perform the same process of reasoning and calculation to make it clear why the maximum negative voltage between point B and ground is -2 volts and why the full cycle of resultant output is as shown. Thus, although the resultant output is the arithmetical difference between the instantaneous d-c voltages across R_1 and R_2 , it is equal in magnitude to the sum of the a-c audio components of the voltages across R_1 and R_2 .

Frequency Characteristic. At the transmitter, a variation in the amplitude of the audio signal produces a proportional frequency deviation of the modulated carrier. At the discriminator of the FM receiver, a deviation in frequency produces a proportional variation in the amplitude of the discriminator output signal. The FM signal is thus reconverted at the receiver into the original audio signal.

To reproduce the original audio signal, the frequency as well as the amplitude characteristic must be reproduced. We have just seen how the amplitude characteristic is restored. What about the frequency characteristic?

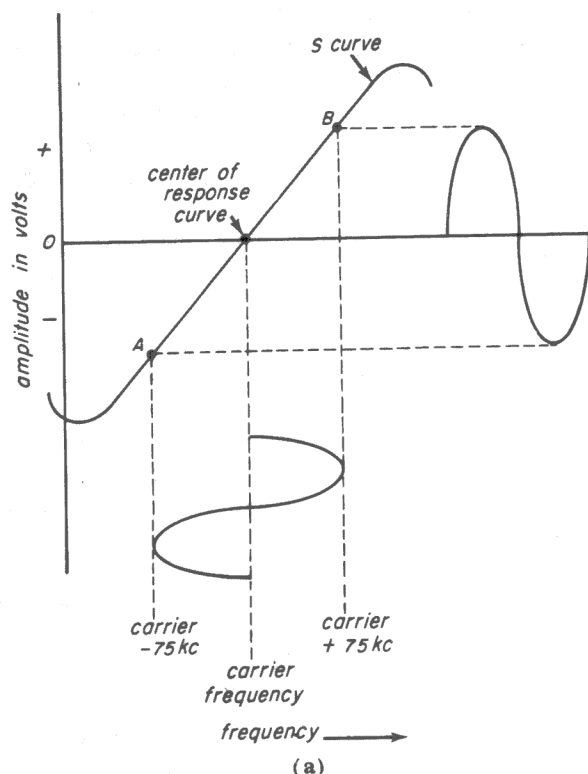
No special circuit is needed to restore this characteristic, because it is never lost or changed. The rate of change of the transmitted FM modulated carrier - that is, the rate at which the frequency deviation produced by the audio signal occurs - is the same as the rate of change or frequency of the original audio signal.

The rate of change of the instantaneous d-c voltages at the output of the discriminator is the same as the rate of deviation of the frequency-modulated i-f carrier at the input of the discriminator.

If a 100-cps audio signal is produced at the transmitter, the FM modulated carrier will deviate back and forth through the center frequency at a 100-cps rate; the modulated i-f carrier at the input to the discriminator will deviate at this same 100-cps rate, and the voltage at the output of the discriminator will change in amplitude at this same 100-cps rate.

The Discriminator Response Curve. We have already mentioned the S curve, which is the response curve of the discriminator. If the section of the curve (Fig. 37-26a) between point A and B is straight, and the FM i-f carrier signal comes in at the center region of the response curve, the audio voltage output of the discriminator will be undistorted.

If the i-f signal comes in at some point other than the center region (Fig. 37-26b)



some of the frequency deviations of the incoming signal will not produce proportional amplitude variations, and audio distortion will result. That is, some frequency deviations will be reproduced along the non-linear portion of the S curve, making the audio voltage output nonlinear or disproportionate (distorted). This situation can occur when the tuning dial of the receiver is not properly set. As a result, the oscillator is not properly tuned and the i-f that results in the converter is not the i-f to which the set is tuned. This condition can be corrected by adjusting the tuning dial of the receiver.

On the other hand, if either the discriminator transformer primary or any of the circuits preceding the discriminator is misaligned, the response curve of the discriminator is itself shifted off center with respect to its zero axis. In Fig. 37-26c, the response curve indicates that the secondary of the discriminator transformer is properly tuned to the i-f carrier frequency but the primary and circuits preceding the primary are not aligned with it.

Effect of Amplitude Modulation on Discriminator. The phase-shift discriminator is

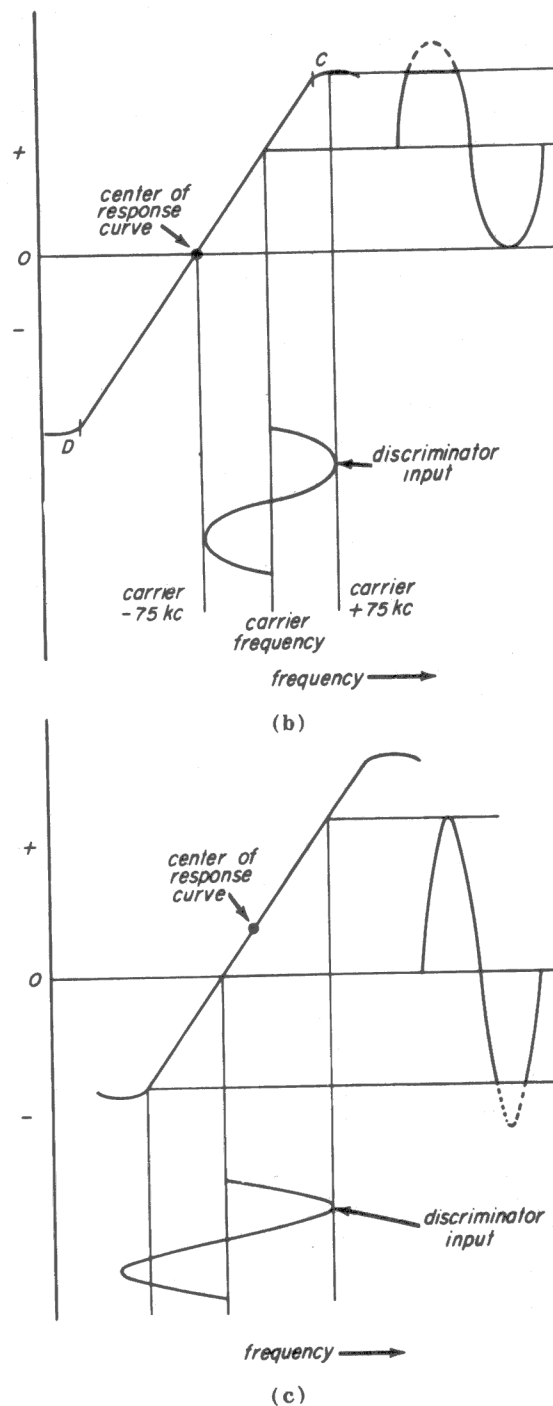


Fig. 37-26

capable of completely rejecting unwanted amplitude modulation that affects incoming i-f signal at the carrier frequency. However, it only partially rejects amplitude modulation that affects signals above and below the carrier frequency. Unwanted amplitude modulation means any rapid change in the amplitude of the frequency-modulated i-f signal coming into the dis-

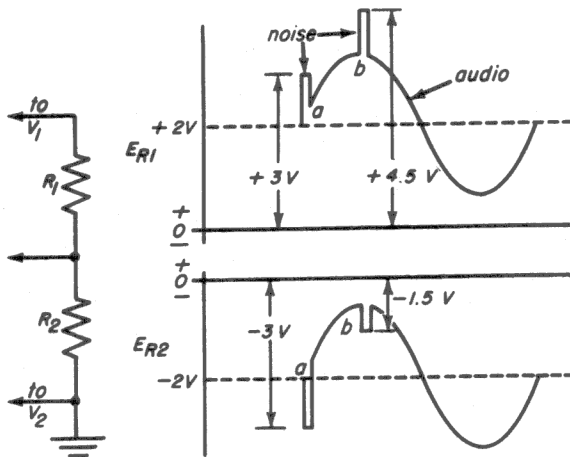


Fig. 37-27

criminator. Noise, rapid fading, or beating of the FM signal with interfering signals changes the amplitude of the FM signal.

Unwanted amplitude modulation of the FM signal appears as opposing variations of the output voltages across the discriminator load resistors. In Fig. 37-27 a noise pulse appears in the voltage E_{R1} across R_1 at time a and in the voltage E_{R2} across R_2 at the same time. At this instant, the FM input signal is at the carrier frequency and all output voltages across the resistors are instantaneously equal and opposite. Since the instantaneous d-c voltages across the resistors are equal and opposite, the audio voltages are equal and opposite, and the noise voltage is equal and opposite, the resultant output voltage ($E_{R1} - E_{R2}$) at this instant (a) is zero — there is no noise pulse in the resultant output.

At another instant, labeled b in the graphs, a noise pulse of the same input amplitude as before appears in the output voltage of the discriminator. At this instant, the FM signal is not at carrier frequency. As a result, the instantaneous d-c voltage, the instantaneous audio voltage, and the noise pulse voltage are all larger across one resistor (R_1) than across the other resistor (R_2). None of these voltages is completely canceled in the resultant difference voltage (Fig. 37-28). The noise pulse appears as an undesired amplitude variation in the graph of the resultant output voltage. The fact that the d-c and audio voltages do not cancel completely in

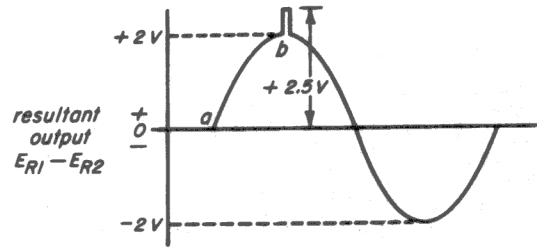


Fig. 37-28

the resultant output is, of course, part of the normal operation of the circuit and has no undesired effect.

Since the discriminator can be adversely affected by amplitude variation, the variations must be eliminated in a limiter, a stage that precedes the discriminator. By eliminating amplitude variations, such as those produced by noise, fading, etc., the limiter prevents such variations from reaching the input to the discriminator.

De-emphasis Circuit. A de-emphasis circuit is an RC filter (Fig. 37-29) generally used at the output of a discriminator or any other FM detector used in the receiver to filter out high-frequency noise. Detected noise tends to be most troublesome at the higher audio frequencies. The resistor and capacitor in a standard de-emphasis network often have a time constant of 75 microseconds.

The de-emphasis circuit also restores the audio frequency spread of signals to the levels they would have had if pre-emphasis had not been applied to them at the transmitter. Pre-emphasis makes the high-frequency audio signals larger in amplitude; the high-frequency attenuation that takes place

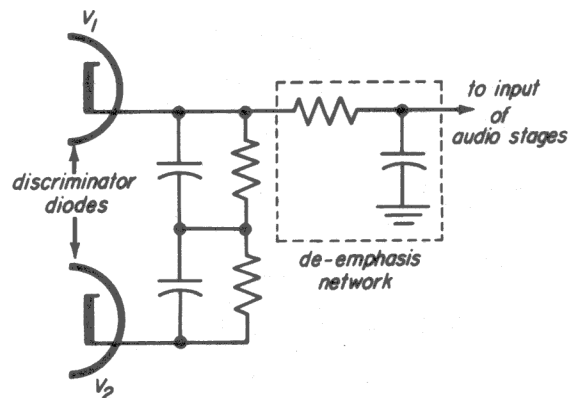


Fig. 37-29

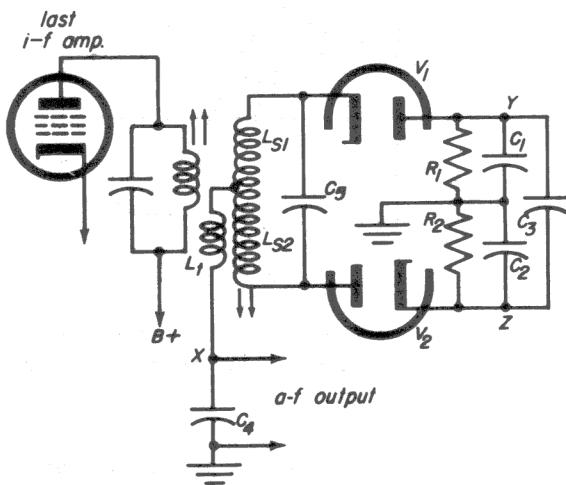


Fig. 37-30

in the de-emphasis circuit restores the signals to their original amplitudes and at the same time reduces noise picked up after the signal has left the transmitter.

37-4. THE RATIO DETECTOR

Used without a limiter, the Foster-Seeley discriminator responds to some amplitude variations in the incoming signal. But the ratio detector does not require a limiter, because it is not sensitive to amplitude modulation.

Basic Ratio Detector-Circuit. The circuit of a ratio detector is illustrated in Fig. 37-30. The transformer used is somewhat different from the one employed in a discriminator stage. In addition to the tuned primary and secondary, it has a small winding, L_t , known as a *tertiary winding*. This L_t winding has the same function as the coupling capacitor connected between the upper end of the

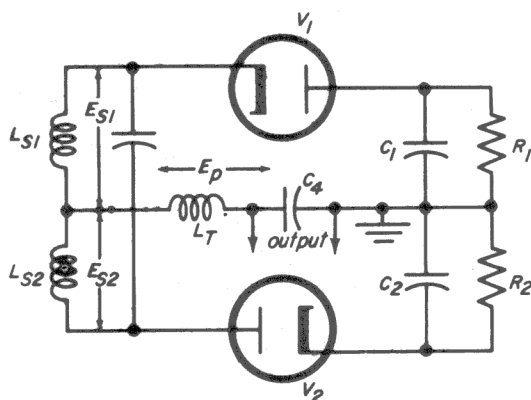


Fig. 37-31

primary and the center tap of the secondary in a discriminator-type transformer. That is, it causes a voltage that is in phase with the primary voltage to be developed between the center tap of the transformer and ground.

The network C_3 , R_1 , and R_2 in the output circuit of the diodes form a network whose time constant is approximately 0.2 seconds. Later we will discuss how this network makes the ratio detector substantially immune to amplitude variations. The audio voltage output of the ratio detector in Fig. 37-30 is developed between point X and ground.

How the Ratio Detector Works. The input voltages applied to the ratio detector diodes are similar to the input voltages at the diodes of the Foster-Seeley discriminator. The simplified diagram shown in Fig. 37-31 illustrates the relationship of these voltages.

Since the reactance of C_4 is relatively low at the i.f., the primary voltage E_p is effectively applied across L_t to the center-tap of of the secondary and ground. The voltage that is applied to the input of the upper diode circuit is the vector sum of E_{s1} and E_p . Similarly, the voltage applied to the input of the lower diode circuit is the vector sum of E_{s2} and E_p .

The vector relationships of these voltages when the incoming signal is at, below and above the carrier frequency are the same as in the case of the Foster-Seeley discriminator as shown in Fig. 37-32. E_{v1} and E_{v2} are the resultant input voltages to V_1 and V_2 , respectively, and are equal at the carrier frequency, (a); when the incoming signal is above the carrier (b) E_{v1} exceeds E_{v2} ; when the incoming signal is below the carrier (c), E_{v2} exceeds E_{v1} .

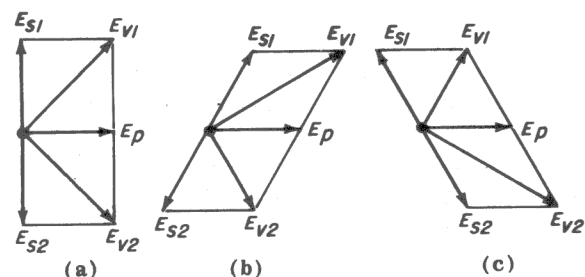


Fig. 37-32

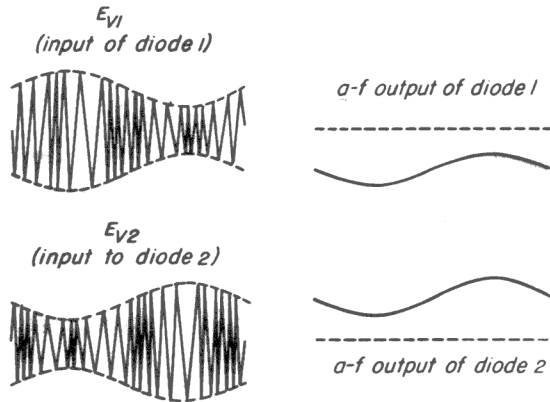


Fig. 37-33

E_{v1} and E_{v2} , the resultant input voltages to diodes V_1 and V_2 , vary in frequency and amplitude, as indicated in Fig. 37-33. The variations in amplitude occur at an audio rate. The intermediate-frequency variations of the input signal are filtered out at the ratio detector output by R_1C_1 , R_2C_2 , and C_4 (Fig. 37-34). Only the audio frequency amplitude variations encounter enough reactance in the ratio detector output circuit to build up appreciable voltages. Fig. 37-33 shows the audio output voltage wave form of each diode with respect to point X in Fig. 37-34. The audio output voltage is taken off across C_4 . Since C_4 is, in effect, a short-circuit at intermediate frequencies, only audio signals are able to build up voltages across it.

The currents flowing through V_1 and V_2 — I_1 and I_2 respectively — pass in opposite directions through L_t and C_4 (Fig. 37-34). The net current in C_4 is equal to the differ-

ence in amplitudes of the opposing currents. The amplitude of current I_1 at any instant depends on the value of E_{v1} at that instant; the amplitude of current I_2 depends on the value of E_{v2} .

At the carrier frequency, E_{v1} and E_{v2} are equal. Thus, at the instant that the carrier is not modulated, the instantaneous currents flowing through V_1 and V_2 are equal. The currents of V_1 and V_2 flow through L_t and C_4 in opposite directions. This means that the net instantaneous d-c and audio current flowing through L_t and C_4 is zero when the signal is at carrier frequency. Since the audio output (voltage) is taken off across C_4 , a zero current flow through C_4 means no audio output at this time.

At any instant when E_{v1} and E_{v2} are unequal, the instantaneous audio and d-c voltages across C_4 have the same polarity as the greater voltage. Just as in the discriminator, the variations of the instantaneous d-c voltage output follow the FM modulation of the carrier. When the instantaneous voltage outputs are plotted, the result is an a-f and a d-c output. However, the audio output across C_4 is half what the output would be if the circuit were connected as a discriminator. The other half of the audio develops across C_1 and C_2 .

Action of AM Rejection-Capacitor C_3 . Capacitor C_3 (Fig. 37-30) is essential for AM rejection. Let's consider what would happen if C_3 were not used. If the amplitude of the FM carrier tended to rise, the amplitude of the incoming signal voltages applied to diodes V_1 and V_2 would rise. As a result, the diode currents would rise, and the larger difference in the currents would cause a larger voltage variation to be developed across the audio output load. This variation would reproduce the undesired amplitude modulation signal.

Ideally, the incoming i.f. should be a frequency modulated input without any amplitude variation. It is the job of the ratio detector to convert the frequency modulation to amplitude modulation at the input of each diode. But actually, the incoming i.f. may vary in amplitude before it

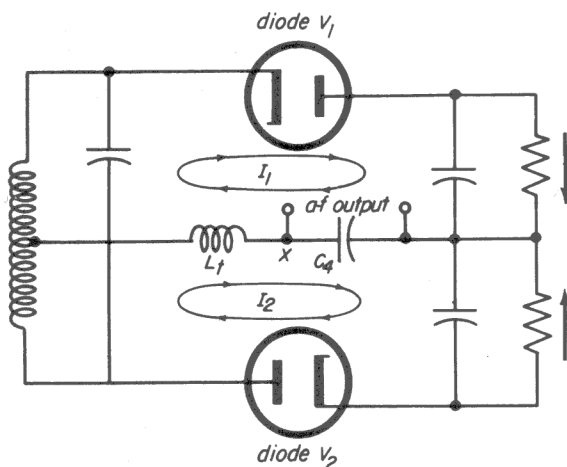


Fig. 37-34

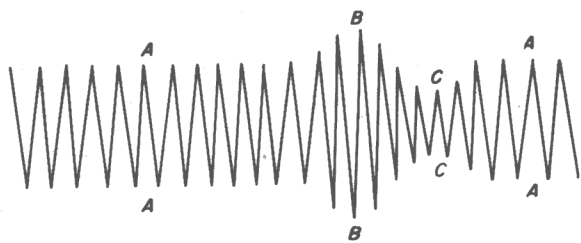


Fig. 37-35

reaches the diodes due to what happens in the r-f and i-f stages. A strong signal may cause either the i-f or r-f stages to operate in a nonlinear fashion. As a result, the outputs of these stages may suffer distortion. Distortion, as you have learned in earlier lessons, actually means that the output of a stage is made up of unwanted harmonics in addition to the desired signal. The unwanted harmonics beat with the desired FM to produce unwanted AM. Usually the amount of distortion is greater when the incoming FM carrier has a high amplitude (that is, when the receiver is tuned to a strong station). A strong incoming signal causes a tube to operate over a larger part of its response curve. If the signal is strong enough, it will be too much for the linear portion of the response curve of the tube and cause the tube to operate partially on the nonlinear portion of its response curve. This phenomenon is called *predetection*. This type of amplitude modulation, as well as other amplitude modulation of relatively long duration, such as the modulation that might be caused by 120-cycle a-c (hum) voltage from the receiver power supply is the type of amplitude modulation that the AM rejection action of a ratio detector is designed to handle. It also handles medium-amplitude short-duration pulses. Do not confuse this type of AM with the momentary AM produced by brief high-amplitude noise pulses like those radiated from automobile ignition systems. A brief strong noise pulse is incompletely rejected by action of C_3 and handled by a balancing effect similar to the balancing effect that occurs in the discriminator.

Capacitor C_3 is the component that provides the unique AM rejection property of the ratio detector shown in Fig. 37-30.

The value of C_3 is chosen so that a relatively long time will be taken for C_3 to

charge and discharge. This time is so long that it is considerably longer even than the longest AM variations. In this respect, C_3 is different from C_1 and C_2 , which do respond to AM variations.

The effect of C_3 is partially upon the stage of i-f amplification preceding the ratio detector. Ordinarily, when the i-f signal does not have unwanted audio modulation, the current flowing in the ratio detector, including C_3 , is not high enough to cause an increase in the normal charge on C_3 . Therefore L_t (Fig. 37-30) does not demand an abnormal amount of power from the primary. However, when the unwanted amplitude modulation is strong enough, — that is, when an exceptionally high voltage is applied to C_3 , this voltage is high enough to overcome the normal bucking charge on capacitor C_3 so that C_3 begins to charge. As a result, current flows into capacitor C_3 . When current flows into capacitor C_3 the transformer must deliver an abnormally large amount of power. The effect is to make C_3 , and therefore the entire ratio detector, a low impedance with respect to the primary of the transformer. As you know, when the output of an amplifier tube is applied to a low impedance, the output of the amplifier is reduced. The effect of C_3 is to reduce the excess output of the i-f amplifier to the level it would have if there were no unwanted AM on the i.f.

In Fig. 37-35, you can see the normal FM modulated i.f. (A). The unwanted AM modulation (B) is superimposed on the i.f. It is this unwanted AM modulation that C_3 acts to eliminate. Thus, the effect of C_3 is to make the ratio detector a variable im-

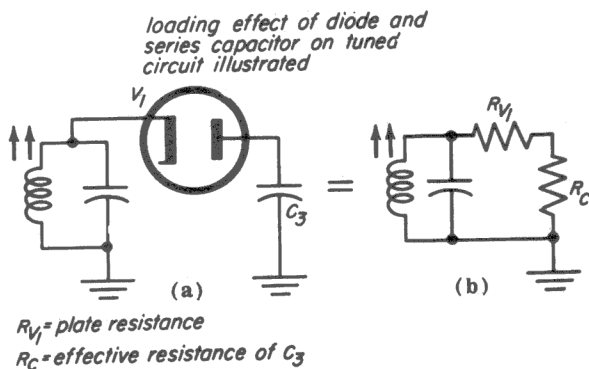


Fig. 37-36

pedance that responds to changes in the amplitude of the input.

So far we have discussed the unwanted AM peaks. But the unwanted AM also causes the FM to drop below its desired level (C). C_3 acts to fill up these low-level intervals.

Normally, C_3 is charged to the voltage level of the i-f carrier. When the rectified i-f voltage applied across C_3 falls below the level to which C_3 is charged, C_3 discharges across the load into C_1 and C_2 . Thus, the voltage across the ratio-detector load capacitors (C_1 and C_2) is kept up to the level that would prevail if there were no unwanted modulation depressing the normal amplitude level of the FM carrier.

Operation of the circuit to reject unwanted AM can be further described in terms of the Q of the transformer secondary circuit. As shown in Fig. 37-36a, the tuned secondary circuit is loaded by the current (or charge) taken through the diodes by C_3 . The loading effects of the diodes and C_3 is equivalent to placing two resistors in series across the tuned circuits as shown in b. One of the resistors is variable. The smaller the value of the variable resistor, the more heavily is the tuned circuit loaded, the lower the Q , and the smaller the gain of the tuned circuit; the higher the Q , the greater is the gain of the tuned circuit. This is in addition to the effect (already explained) of varying the load on the transformer primary winding.

The effective resistance offered by C_3 is equal to the voltage across it (E_c) divided by the current flowing through it (I_c), or:

$$R_c = \frac{E_c}{I_c}$$

Since E_c remains relatively fixed, due to the long time constant of C_3 , whenever I_c increases during a momentary increase in carrier level R_c goes down and the loading of the tuned circuit becomes heavier, lowering the Q and the detector sensitivity. Whenever I_c decreases as the carrier returns to its normal level, R_c goes up, the

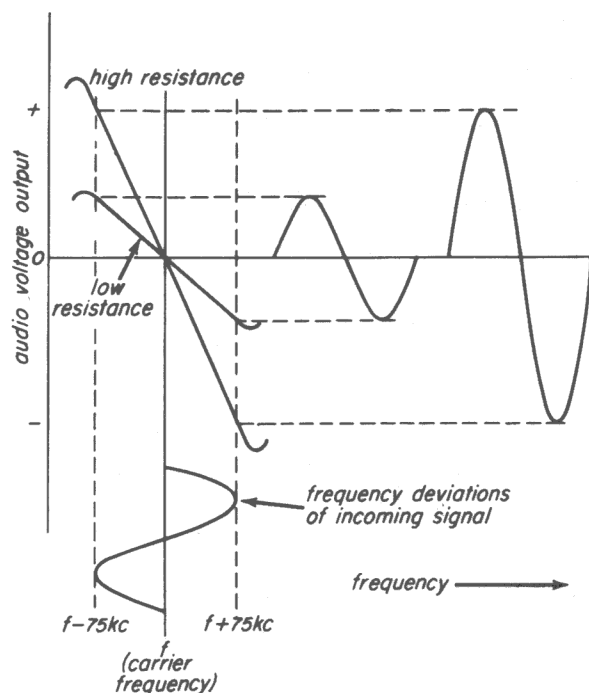


Fig. 37-37

Q of the tuned circuit rises, increasing the detector sensitivity.

I_c goes up when the incoming signal starts to rise in amplitude; I_c goes down when the incoming signal starts to fall in amplitude.

The effect of changes in the effective resistance offered by C_3 on detector sensitivity is illustrated in Fig. 37-37. Note that the lower the resistance, the smaller is the voltage output for the same amount of frequency deviation at the detector input.

Ratio Detector Output. The ratio detector circuit shown in Fig. 37-38a is the same as the circuit in Fig. 37-30, 31, and 34, except that a ground connection has been moved from where it was located in earlier figures. In Fig. 37-38a the old position of the ground is shown in dashed line. As shown in solid line, its new position is on the other side of C_4 . It is possible to shift this ground without altering the way in which the circuit operates because C_4 is a short circuit to i-f voltage. C_1 , C_2 , and L_t are similarly returned to ground no matter which side of C_4 the connection is on. The output is still taken across C_4 . The only effect of shifting the

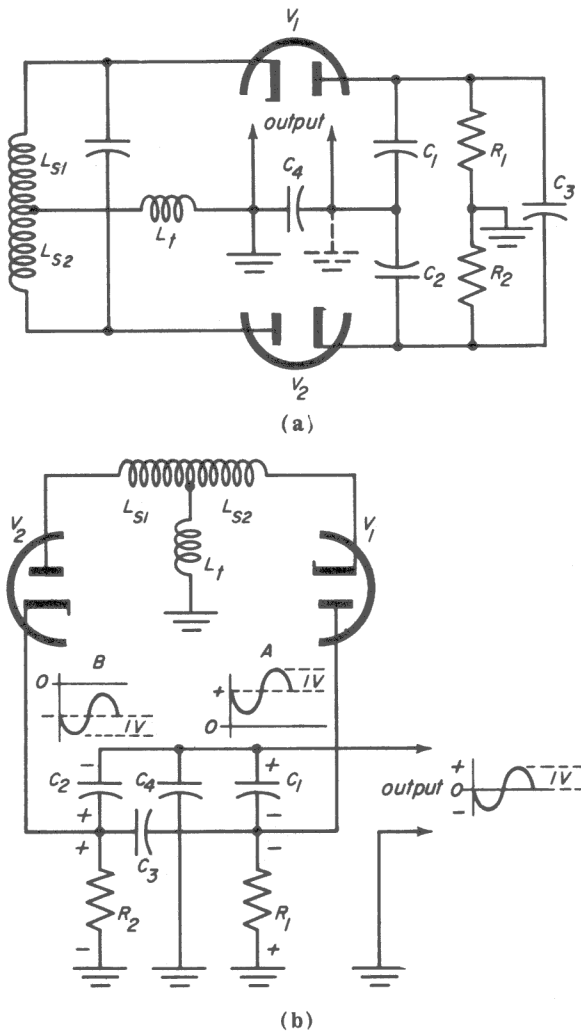


Fig. 37-38

ground is to invert the polarity across C_4 of the varying d-c output of the diodes (shown in Fig. 37-33). The audio variations of the d-c output voltage are still in phase. When the balanced ratio detector circuit of Fig. 37-38a is redrawn as in Fig. 37-38b it is easier to see how the output is taken across two capacitors in parallel. The effect of this output arrangement is different from the output arrangement used in the discriminator. In the discriminator the output is the difference between the outputs of opposite polarity across two load resistors in series. In the balanced ratio detector, the output is of opposite polarity across two capacitors (C_1 and C_2) in parallel. Notice the polarity of capacitors C_1 and C_2 . The fact that they are of opposite polarity when the ratio detector is operating means that the d-c component in the output of one capacitor is equal in amplitude and opposite in polarity to the d-c component of the other

capacitor, as shown at A and B of Fig. 37-38b. The a-c components of the two capacitors are in phase. In calculating the output of the capacitors, we first cancel out the equal and opposite d-c components. This has the effect of shifting both a-c components to the zero axis. Since the two a-c components are in phase and in parallel, this amounts to superimposing the a-c component of one capacitor right on top of the a-c component of the other capacitor. Thus, all you can see on the graph (C) is the output across one of the two capacitors (Fig. 37-38b). This is actually what the resultant output of the ratio detector is—the output across one of the two capacitors (C_1 or C_2). Graph C shows the resultant output. The resultant output voltage appears across C_4 which is in parallel with both C_1 and C_2 .

We have just illustrated graphically the rule that when two voltage sources of equal internal impedance are applied in parallel to an impedance, the total voltage is equal to the sum of the voltages from each of the sources divided by the number of voltage sources.

In A and B of Fig. 37-38b, there are two a-c voltages, each of which has a peak amplitude of 1 volt. The total from both voltage sources is 2 volts. Dividing the 2 volts total by 2, the number of voltage sources, we get a total of 1 volt, which is the same result we got when we used the graphic method of superimposing one a-c component upon the other. The superimposing technique works only when the voltages are equal. However, this means that it always works when you analyze the output of a balanced ratio detector. Balanced ratio detectors always have an output, the a-c components of which are in-phase and equal in amplitude. Therefore, you can use the simplified rule that the total output of a ratio detector is equal to the output across one of the diode load capacitors.

The ratio detector gets its name from the fact that the ratio between the instantaneous audio output of one diode rectifier and the instantaneous audio output of the other diode rectifier is constant in the ratio 1:1. This is the same as saying that the two audio

outputs are equal and in phase. The total output of both diodes is held constant by the control voltage. Thus, if the control voltage (approximately equal to the voltage of the i-f input) is 6 volts, for example, the instantaneous d-c output of each diode might be 3 volts, or the instantaneous d-c output of one diode might be 4 volts and that of the other 2 volts, or the instantaneous input of one diode might be 5 volts and that of the other diode 1 volt. The control voltage changes when you tune the receiver from one station to another. But when you have changed stations, the ratio-detection principle applies as before.

Unbalanced Ratio Detector. The *balanced* ratio detector is called balanced because the impedances in series with each diode are equal. Another form of ratio detector is known as an *unbalanced* ratio detector (Fig. 37-39). The diodes are shown enclosed in the same tube envelope; in the case of the balanced ratio detector, the diodes were shown in separate tube envelopes in order to make the circuit easier to understand.

The unbalanced ratio detector uses fewer components than the balanced type. Otherwise, it is similar in its action to the balanced type of ratio detector. Diode signal currents flow in opposite directions through the path consisting of L_t and C_4 , depending upon which diode is conducting, just as in the balanced ratio detector. Due to lack of balancing action, some noise pulses are not cancelled.

Advantages of Receiver Using Ratio Detector over Receiver Using Limiter-Discriminator. A receiver using a ratio detector provides the certain features not provided by a receiver that uses a limiter-dis-

criminator combination. The ratio detector receiver provides quieter over-all tuning because it is quiet at between-station settings; the limiter-discriminator receiver is noisy, because the limiter acts as an amplifier when no station signal is coming in and amplifies the noise present at its input.

A ratio detector receiver retains its full selectivity in the presence of strong signals. In the limiter-discriminator receiver, strong signals cause considerable grid current flow in the limiter, resulting in the loading of the tuned grid circuit as well as the tuned plate circuit of the preceding stage, to which the grid circuit is coupled. The heavy loading lowers the Q of the tuned circuits, broadening their bandpass and reducing their selectivity to the point where undesired signals may possibly get through.

The ratio detector receiver is easier to tune than a limiter-discriminator receiver.

The set owner is accustomed to tuning his AM set to produce loudest volume. Therefore, he tends to tune his FM set in the same way. In an FM set using a ratio detector, the correct setting of the receiver's fine tuning control is the one at which the volume is maximum. Normally no distortion is audible at maximum volume.

In the case of the limiter-discriminator receiver, the maximum volume setting is not the correct setting; only maximum distortion is audible at maximum volume. The set must be tuned for no distortion, not for maximum volume.

The ratio detector receiver is less subject to cross-talk and adjacent-channel interference. In a discriminator-limiter set, considerable grid current flows in the limiter on strong signals. A tube that operates in this way is capable of acting as a mixer or a detector. An undesired signal at the input of the limiter, due to insufficient selectivity of the preceding stages, will beat with the desired signal. Cross-modulation — the audio modulation of the undesired station modulating the desired station — will result. Both audio signals will be heard.

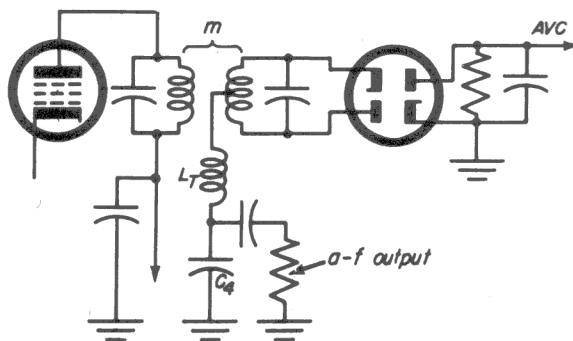


Fig. 37-39

37-5. THE GATED BEAM DISCRIMINATOR

Another type of detector used in FM receivers is known as the *gated beam discriminator*. This type of circuit requires a special type of tube, such as the 6BN6.

The gated beam discriminator requires no limiter, since it has adequate inherent limiting in itself. Also, a first audio amplifier tube may not be required since it may have enough a-f output to drive the power amplifier.

The way in which the gated beam discriminator works can be described in simple terms. At the carrier frequency, the *average* plate current has a certain amplitude. When the frequency of the incoming signal rises above the carrier frequency, the average plate current drops in proportion to the frequency deviation; when the frequency of the incoming signal falls below the carrier frequency, the *average* plate current rises proportionately. Thus, the average plate current varies with the frequency of the incoming signal. The rate of frequency deviation is converted into a varying plate current whose frequency is equal to the rate of deviation. This satisfies one of the fundamental requirements of an FM detection system. Before we discuss the operation of the gated beam discriminator in greater detail, let us first discuss its construction.

The gated beam tube is constructed so that a few volts change on either grid will sharply vary the current from cutoff to saturation. In other words, the e_c-i_b characteristic curve is steep and short, as in Fig. 37-40. This characteristic is accomplished in the following manner.

A sharply-focused beam of electrons is passed through a narrow aperture in a solid positive electrode and is immediately afterward directed against a control grid which is followed by a positive anode. As a result an unusually rapid swing in plate current from cutoff to saturation becomes possible. To produce the desired focusing of the beam of electrons, the 6BN6 uses an electron lens system. An electron lens is similar to an optical lens. An optical lens

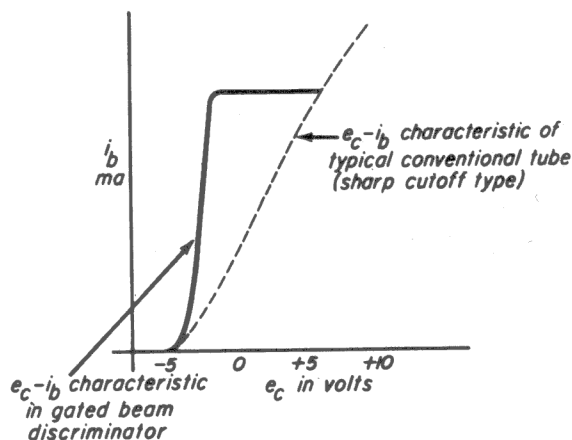


Fig. 37-40

converges light (brings light rays to a focus); also, a lens can also cause light rays to *diverge* (spread apart). An electron lens does the same thing to a beam of electrons. It does it electronically, however; in the gated beam tube, this action is achieved by means of the electrostatic fields developed between sets of electrodes whose structure and potentials are such as to provide the desired shaping of the beam.

Structure of the 6BN6. A cross section drawing of the 6BN6 is shown in Fig. 37-41. The tube contains a filament (not shown in the sketch); a cathode; two control grids, known as the limiter and quadrature grids, respectively; a focus electrode; shield elements; screen grid; accelerator anode; lens electrode; and a plate.

The focus, lens, and shield electrodes are internally connected to the cathode. The accelerator and plate are connected to positive potentials.

The focus electrode, in conjunction with the aperture in the accelerator electrode,

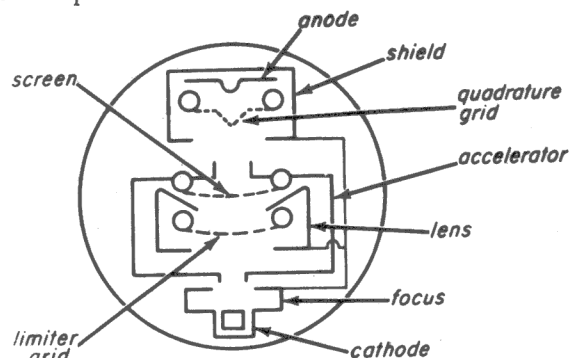


Fig. 37-41

forms an electron gun, similar to the one in a cathode-ray tube. The effect of the gun is to focus a stream of electrons as a thin sheet, on the limiter grid.

After passing through the limiter grid, the electron beam diverges. It is then focused once more by the screen and lens electrodes, and projected through the second and final aperture in the accelerator anode onto the quadrature grid. After passing through the quadrature grid, the electrons in the beam flow to the plate.

Gating Action of 6BN6. Throughout the preceding discussion, we have assumed that the voltage between each control grid and cathode is such that electrons are able to reach the plate. However, let us stress that if *either* grid is more than a few volts negative with respect to cathode, plate current is cut off, regardless of the potential of the other grid. Each grid thus acts as a *gate* on the electron beam.

If the limiter grid is more than a few volts negative, the electron beam is unable to pass it. The beam is, in this case, repelled by the limiter grid and electrons flow back in the direction of the cathode (Fig. 37-42). They are prevented from reaching the cathode by the positive-potential accelerator electrode. Also, because of the accelerator electrode having a small opening, the opening becomes crowded with electrons very rapidly, causing other on-coming electrons to be repelled.

If the limiter grid voltage is less negative than cutoff, the electron beam passes through this grid and continues on to the quadrature grid.

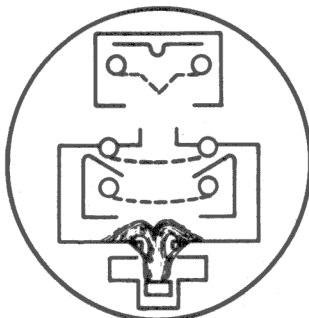


Fig. 37-42

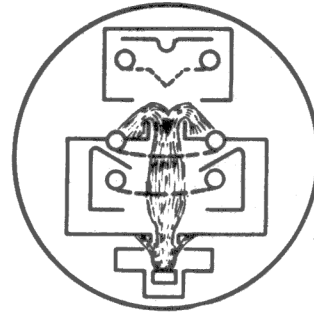


Fig. 37-43

If the quadrature grid voltage is more negative than cutoff, the beam is repelled back from the quadrature grid. The backward-moving beam diverges, and is attracted to the positive accelerator (Fig. 37-43).

If the quadrature grid voltage is less negative than cutoff, the beam (actually a portion of the beam) flows on to the plate.

Accelerator current always flows in the 6BN6, even when the voltage of one or both grids is beyond cutoff and plate current flow becomes zero. The accelerator current forms a major portion of the total cathode current. When plate current is cut off by the negative voltage at one or both grids, the electrons that would have flowed through the plate circuit if the tube had not been cut off flow through the accelerator circuit instead, increasing the accelerator current. As a result, the cathode current is constant.

Limiting Action of Gated Beam Tube.

Let's consider the limiting action of the gated beam tube in greater detail

Referring to the schematic diagram in Fig. 37-44, the incoming i-f signal is applied between limiter grid and ground. The positive and negative swings of the signal drive the plate current from saturation to cutoff, respectively, causing a square wave pulse of current to be produced for each cycle of input signal. This kind of current waveform is needed to minimize the effect on plate current of the undesired amplitude variations of the input signal. Other amplitude variations (due to noise or to inter-

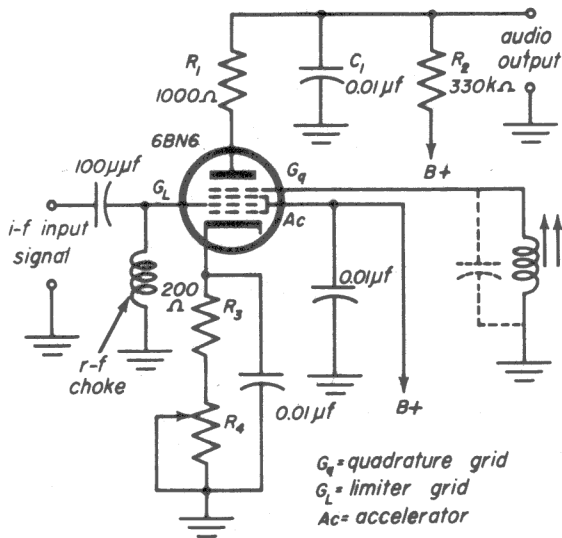


Fig. 37-44

terfering signals) are automatically eliminated when the peaks of the input signal are clipped off (Fig. 37-45).

The bias for 6BN6 is self-biased by cathode current flowing through cathode resistor (R_4 , Fig. 37-44). When R_4 is properly adjusted, the operating point of the tube will be midway between plate current saturation and plate-current cutoff. The design of the tube is such that a cathode bias correct for the operation of the limiter section of the tube is also correct for the operation of the quadrature section; a single cathode resistor is thus capable of biasing the control grids properly.

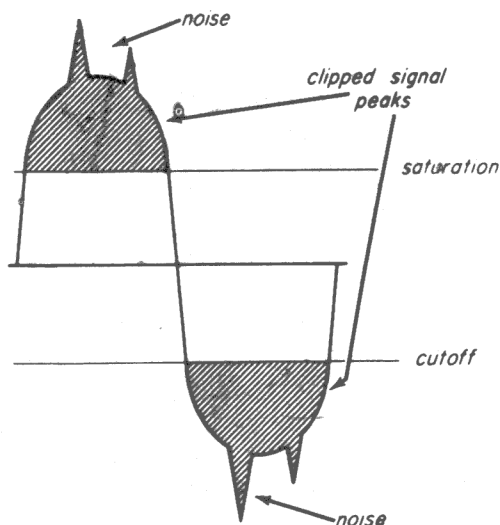


Fig. 37-45

FM Detection Action of 6BN6. The detection of FM signals in the 6BN6 is achieved with the aid of the quadrature grid. The quadrature grid is connected to a parallel-resonant circuit made up of a slug-tuned coil that resonates with the stray circuit capacitance (indicated by dash-line capacitor). This circuit is tuned to the i-f carrier frequency. The pulses of beam current produced by the signals at the limiter grid are coupled to the quadrature grid through *space-charge coupling*. This method of coupling is somewhat similar to the coupling that would take place if a capacitor were connected between limiter and quadrature grids. Due to the space-charge coupling, a signal voltage is developed between quadrature grid and ground that is different in phase from the signal voltage present between limiter grid and ground.

When the i-f carrier is coming in, the signal voltage at the quadrature grid lags the limiter grid signal voltage by 90 degrees. When the frequency of the incoming i-f signal is *above* the frequency of the carrier, the phase angle between the quadrature and limiter grid voltage is *less* than 90 degrees. When the frequency of the incoming i-f signal is *below* the frequency of the carrier, the phase angle between the quadrature and limiter grid voltages is *more* than 90 degrees. These phase differences in the voltages at the two grids produce proportional changes in the amplitude of the plate current flowing through 6BN6. A proportional voltage variation across the load resistance of the 6BN6 consequently takes place. Thus, a frequency deviation of the i-f input signal is converted into an amplitude variation in the output circuit of the 6BN6. The details of this process are considered in the paragraphs that follow.

Phase Relations of Limiter and Quadrature Grid Voltage. Let's assume that the i-f carrier appears between limiter grid and ground. Through space-charge coupling, the quadrature grid circuit will oscillate at the i-f carrier frequency. The tuned circuit Fig. 37-46a is resistive at this frequency (since the circuit reactances cancel each

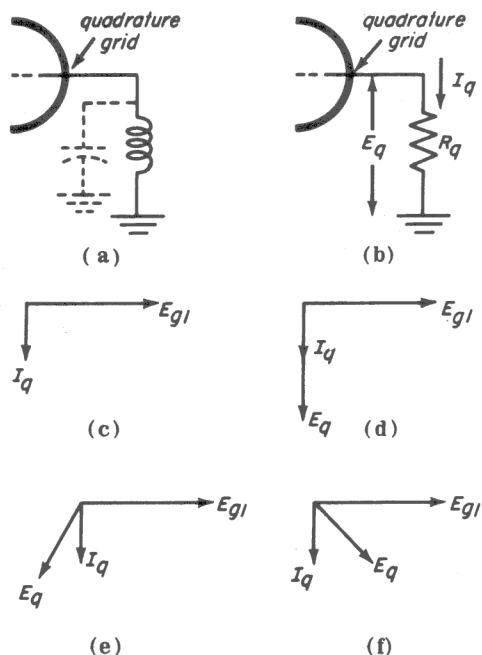


Fig. 37-46

other). The equivalent circuit may therefore be drawn as a resistor (R_q , Fig. 37-46b). Space-charge coupling causes the quadrature grid and the stream of electrons flowing through the 6BN6 to act somewhat (but not exactly) like the two plates of a capacitor. Due to the nature of this coupling, the current through the quadrature grid circuit (I_q) lags the signal voltage at the limiter grid (E_g) by approximately 90 degrees. This phase relationship is indicated in Fig. 37-46c. The voltage developed across R_q is in phase with the I_q current flowing through R_q , since this is always true of the voltage and current relations with respect to a resistor. Therefore, the voltage developed across R_q (that is, the voltage between quadrature grid and ground at resonance, or E_q) lags E_g by 90 degrees (Fig. 37-46d).

This is the condition when the i-f carrier is at the center frequency. When the frequency of the incoming signal rises above the carrier, the quadrature-grid parallel-resonant circuit is no longer resistive. The reactance of the capacitor is now less than the reactance of the coil. A larger signal current will therefore flow through the capacitor than through the coil; the net current (I_q) will therefore be capacitive. Therefore, I_q will lead E_g the voltage developed across the tuned circuit.

Or we can say E_q will lag I_q . Since the phase relation of I_q and E_g remains fixed at 90 degrees, when E_q lags I_q , the phase angle between E_q and E_g increases to more than 90 degrees (Fig. 37-45e).

When the incoming signal drops below the frequency of the i-f carrier, the reactance of the capacitor becomes greater than the reactance of the coil. Therefore, more signal current flows through the coil than through the capacitor, and the parallel tuned circuit will be inductive. Therefore, current I_q lags quadrature grid voltage E_q , or E_q leads I_q . This causes the phase angle between E_q and E_g to become less than 90 degrees (Fig. 37-45f). Note that the phase between E_g and I_q remains at 90 degrees at, above and below center frequencies. However, the phase between E_q and E_g varies with frequency.

Let's use waveforms (Fig. 37-47) to help us analyze the effects of phase differences in the limiter and quadrature grid voltages on the 6BN6 plate current.

At the center i.f., if the signal voltage on the quadrature grid is 90 degrees out of phase with the limiter grid signal voltage, the waveforms present are as indicated in Fig. 37-47a. Plate current is cut off by the negative voltage peak at one grid during time T_1 , and by the negative voltage peak at the second grid during time T_2 .

The total time of cutoff is time T_3 . Note that intervals T_1 and T_2 overlap to some extent. If T_1 and T_2 overlapped completely, cutoff time T_3 would be a minimum. If T_1 and T_2 didn't overlap at all, cutoff time T_3 would be a maximum.

You will observe, in Fig. 37-47a, that each plate current pulse begins when the signal voltage at one grid rises above cutoff and ends when the signal voltage at the other grid moves below cutoff.

Above center frequency, the phase angle between the limiter and quadrature grid voltages is greater than 90 degrees. The negative peaks of the quadrature and limiter grid voltages move further apart (Fig. 37-47b). There is a cutoff voltage on one

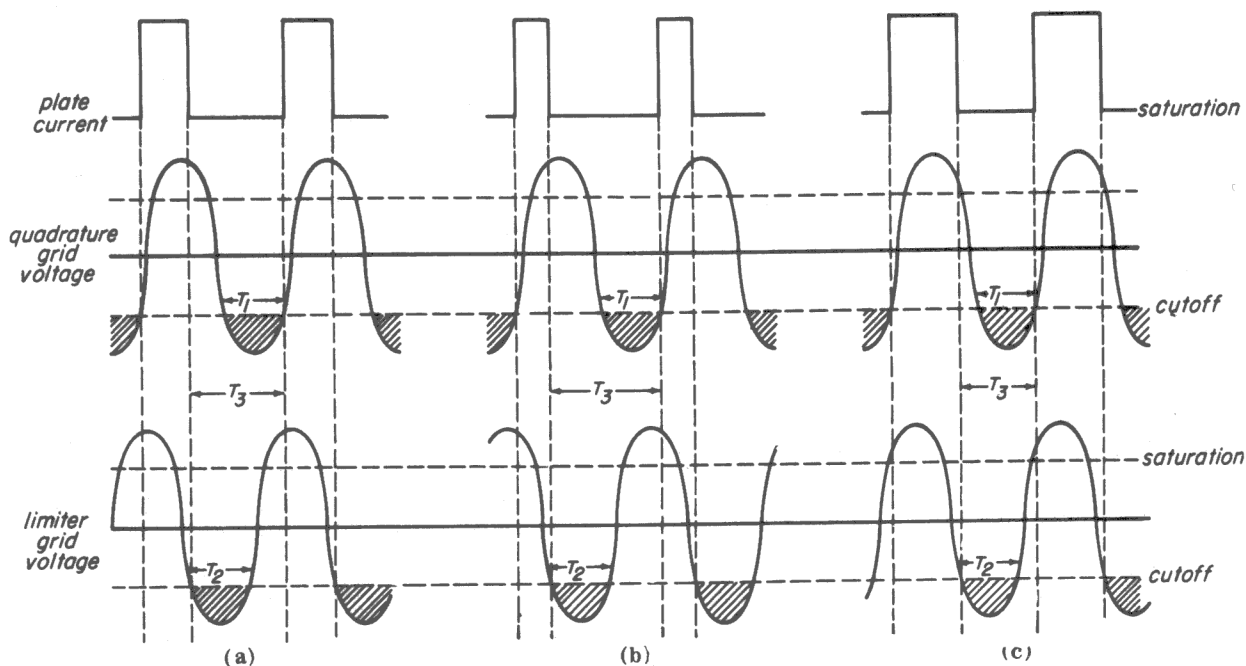


Fig. 37-47

or the other grid over a longer part of the cycle. Plate current flows for a shorter portion of the cycle, and the width of the plate current pulses is reduced.

Below center frequency, the phase angle between quadrature and limiter grid voltages becomes *less* than 90 degrees (Fig. 37-47c). The negative peaks of the quadrature and limiter grid voltages move closer together. There is cutoff voltage to be present for a shorter part of the cycle. Plate current flows for a longer portion of the cycle and the width of the plate current pulses increases. Thus the smaller the phase angle between the voltages at the two grids, the wider the plate current pulse and the higher the average plate current, and vice versa.

We pointed out earlier in the lesson that the phase angle between the limiter and quadrature grid voltages was 90 degrees when the i-f carrier was coming in, more than 90 degrees when the incoming i-f signal was above the center frequency, and less than 90 degrees when the incoming i-f signals was below the carrier.

When the i-f signal goes above the center frequency, and the phase angle between the limiter and quadrature grid voltages rises, the average plate current drops. When the i-f signal goes below the carrier, and the phase angle between the limiter and quadra-

ture grid voltages becomes lower, the average plate current increases.

The amount of rise or fall in average plate current at any instant is proportional to the phase difference between the limiter and quadrature grid voltages, which is, in turn, proportional to the frequency deviation of the incoming signal. The change in plate current is therefore proportional to frequency deviation. Frequency variations in the incoming i-f signal are converted into corresponding amplitude variations in current. Let's consider how the amplitude variations in plate current are converted into audio signals. The plate current pulses occur at an intermediate-frequency repetition rate. The average amplitude of each pulse corresponds to the instantaneous amplitude of the *original audio signal*, as represented by the frequency deviation instantaneously present at the input of the 6BN6.

If a continuous line is drawn connecting the average values of each pulse, the waveform of the original audio signal would be reproduced. This is roughly indicated in Fig. 37-48. The average plate current level of several pulses is represented in Fig. 37-48. Note that the average value of each pulse can be represented by a plot point (Fig. 37-48b), representing plate current. If the plot points are connected together

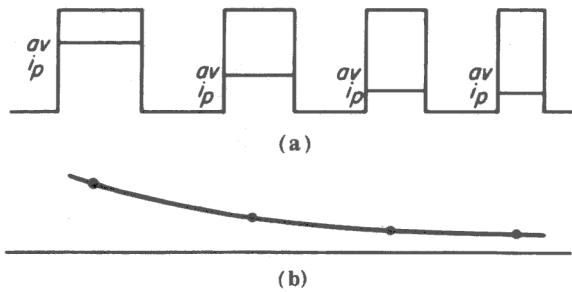


Fig. 37-48

(which, in effect, the filter R_2C_1 does), there will be variations in plate current corresponding to the original modulating signal.

The intermediate-frequency variation must be removed from the plate current, without removing the average current variation, which occurs at an audio-frequency rate. To do this, a low-pass filter is inserted at the plate of the 6BN6. The filter is similar to the one used in a diode detector to remove the i-f signal from the audio signal developed across the diode detector load resistor.

The filter is made up of a resistor and a capacitor (R_2 and C_1 in Fig. 37-49). It removes i-f signals from the plate load of the 6BN6, permitting only audio signals to develop an output voltage in this circuit. The size of capacitor C_1 is such that both proper de-emphasis and i-f filtering are provided.

Insertion of R_1 , as shown in Fig. 37-44, causes some i-f voltage to be developed across R_1 . I-F voltage is developed across R_1 because filtering of the i-f signal current does not occur before the output side of R_2 is reached. The i-f voltage developed across R_1 is coupled back to the quadrature tuned circuit through the interelectrode capacitance present between the plate and the quadrature grid. The phase difference introduced by this capacitance, and the wiring associated with it, is such that the i-f voltage fed back is in phase with the i-f voltage developed across the quadrature coil. The voltage across the quadrature tuned circuit is thus reinforced or strengthened.

At the same time, R_1 widens the band-pass of the quadrature circuit. The inter-

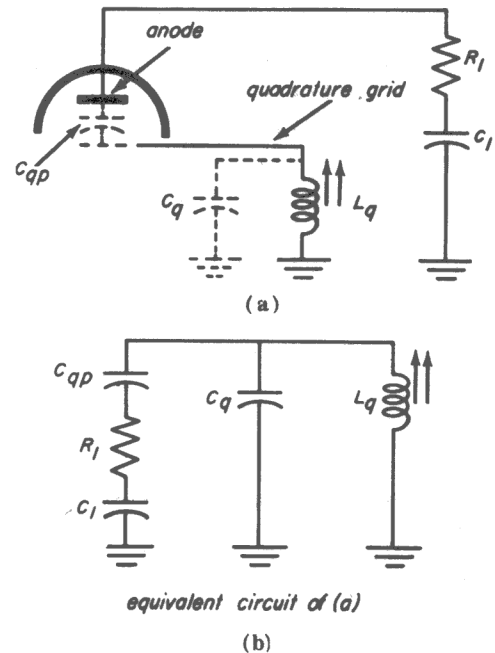


Fig. 37-49

electrode capacitance between the quadrature grid and plate forms a part of the total capacitance in the quadrature tuned circuit (Fig. 37-49). Resistor R_1 is in series with the interelectrode capacitance, C_{qp} , and damps the tuned circuit, that is, lowers its Q and broadens its bandpass. Ordinarily, such damping occurs at the expense of gain. The AM rejection characteristic is also adversely affected. However, linearity and harmonic distortion are improved. Due to the coupling back of i-f signal from plate to quadrature grid, however, the damping effect of R_1 on the quadrature grid is more than compensated for. Typical values of R_1 are from about 300 to 1,000 ohms. Increasing the value of the resistor reduces the harmonic distortion but decreases the audio output and reduces AM rejection, except at low signal levels.

37-6. THE FM LIMITER

The limiter stage in the FM receiver removes amplitude modulation from the FM signal. A limiter is necessary whenever the FM detector used is not able to reject amplitude modulation. The Foster-Seeley discriminator requires a limiter in front of it, since it is highly sensitive to amplitude modulation. The ratio detector is insensitive to AM.

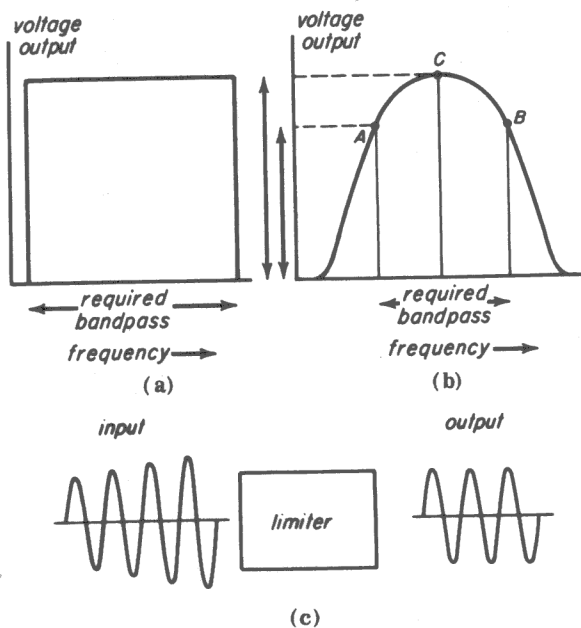


Fig. 37-50

Noise is a source of undesired amplitude modulation. Another source is the amplitude variations in the FM signal introduced by the tuned circuits of the receiver.

Ideally, signals in the i-f passband should all produce equal voltages across the tuned circuits, as shown in Fig. 37-50a. This would require a flat tuned circuit response. But such a response cannot economically be attained. As shown in Fig. 37-50b some rounding off of the response must be tolerated. As a result, FM signals that fall close to the center of a tuned circuit's response (point C) will develop more voltage output in the tuned circuit than signals that fall closer to the ends of the response (points A and B). If such variations in amplitude were not removed, they would produce amplitude variations at the output of the discriminator that were unrelated to the frequency deviations of the FM signal; the result would be distortion of the audio signal.

The limiter clips both peaks of the i-f signal applied to it. The applied signal varies both in frequency and amplitude at the input of the limiter; the signal voltage at the output of the limiter, however, retains only its frequency variations, the amplitude variations are all removed (see Fig. 37-50c).

Typical Limiter Circuit. A typical limiter circuit is illustrated schematically in Fig.

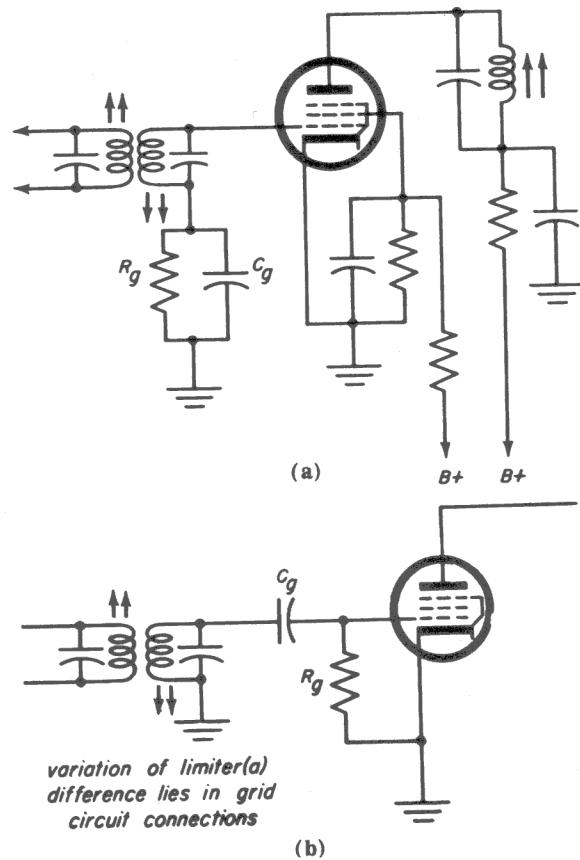


Fig. 37-51

37-51a. A slightly different version of this circuit – the grid circuit is different – is shown in Fig. 37-51b. Limiting takes place in both the grid and plate circuits. Let's consider first the action occurring at the grid. Initially the limiter tube is operated with zero bias. Grid-leak bias is then developed by the incoming signal. The larger the incoming signal, the greater the grid-leak bias, and vice-versa. Let's review this grid-leak bias action. During the first positive half cycle of the input signal, the control grid becomes positive with respect to cathode. Current flows between cathode and grid. The C_1 charges up to some extent (See Fig. 37-52a). It is the voltage across C_1 that is the bias. Looking between control grid and ground there is the bias (grid being made negative with respect to cathode) in series with the signal voltage of the secondary. Thus, the bias is derived from the signal voltage, and is therefore known as signal bias as well as grid leak bias.

During the negative half cycle of input signal voltage, some of the charge built up on C_1 during the preceding half cycle leaks off through R_1 (Fig. 37-52b). C_1 discharges

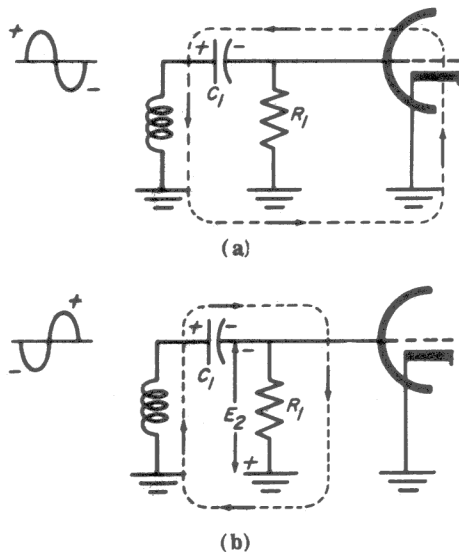


Fig. 37-52

only slightly, because of the relatively long discharge time constant of $R_1 C_1$.

At the time of the next positive half cycle, additional grid current flows. Current starts to flow when the positive value of the signal voltage exceeds the negative voltage across C_1 , charging it to a higher value of voltage. During the following negative half cycle, C_1 discharges through R_1 .

This action continues, with the bias becoming larger and larger. Grid current cannot flow until the positive peak of the grid signal voltage exceeds the negative bias (across C_1). Since the bias is continually increasing, the level of positive signal voltage needed to exceed it gets continually larger. This action continues until C_1 is charged to a voltage equal to the peak value of the signal voltage. The action of developing grid-leak bias can be shown in another way. Looking at Fig. 37-53 charging and discharging curves for $R_1 C_1$ are shown. During intervals of charging (1-2, 3-4, etc.), the voltage across C_1 rises.

The capacitor C_1 absorbs more charge during the interval of charging (grid current flow) than it loses during the interval that grid current is not flowing. The C_1 charging current, however, has been getting smaller and smaller, since an increasing number of electrons accumulates on C_1 . These electrons cannot all leak off quickly enough through R_1 , therefore a larger and larger number of them remain on the grid side of

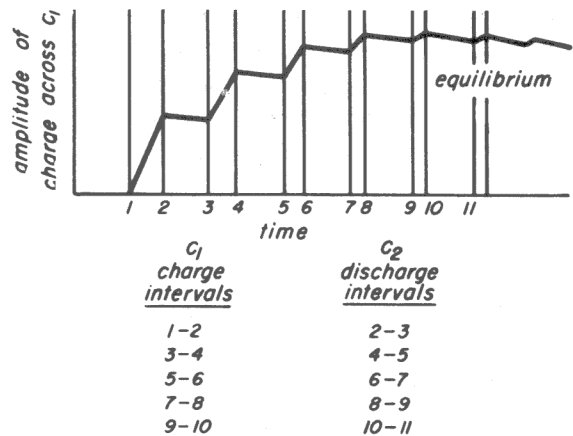


Fig. 37-53

C_1 . A point is reached, where the number of electrons C_1 gained during charge time is equal to the number of electrons lost during its discharge. This condition is referred to as equilibrium, or a state of balance. While we have taken some time to discuss it, equilibrium is actually reached in less than a fraction of a second. Figure 37-54 shows the grid leak bias voltage reaching equilibrium.

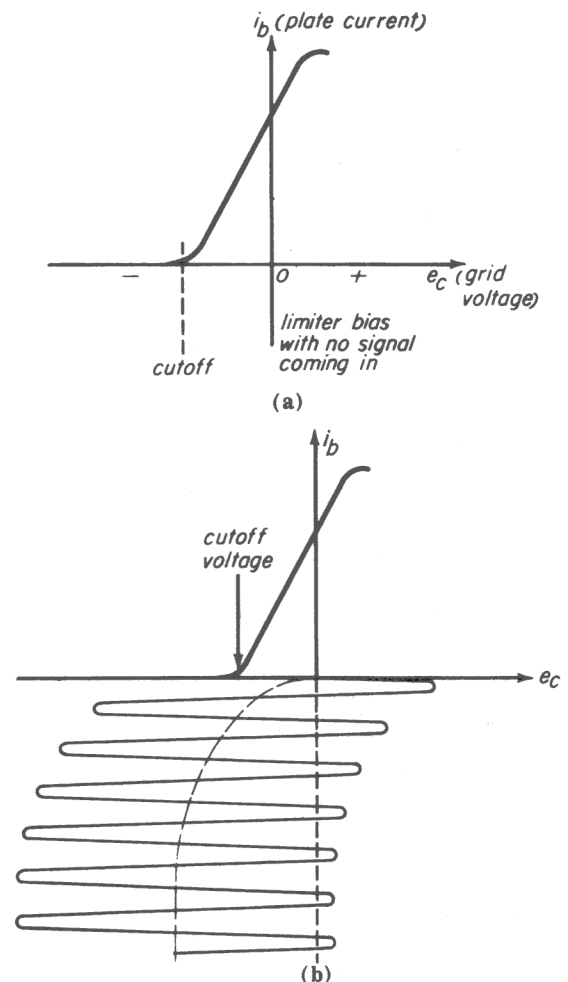


Fig. 37-54

The grid-leak voltage present across C_1 varies only slightly during each cycle from now on, since the charge and discharge currents are relatively small. The grid-leak voltage across C_1 becomes, essentially, a d-c voltage, and biases the limiter.

How Negative Peaks of the Signal Are Clipped. By choosing proper values of R_1 and C_1 , using a tube having suitable characteristics, with the plate and screen grid operated at low d-c voltages. Feeding in a certain minimum of input signal, the limiter grid-leak bias will be at or beyond cut-off for a wide range of incoming signals. As shown in Fig. 37-55 negative peaks of the incoming signals are clipped because the grid-to-cathode voltage of the limiter is driven beyond cutoff by these peaks. No plate current flows during these intervals.

When no plate current flows, no signal voltage variations will be developed at the output of the limiter, even though there is a varying signal voltage at the input. Am-

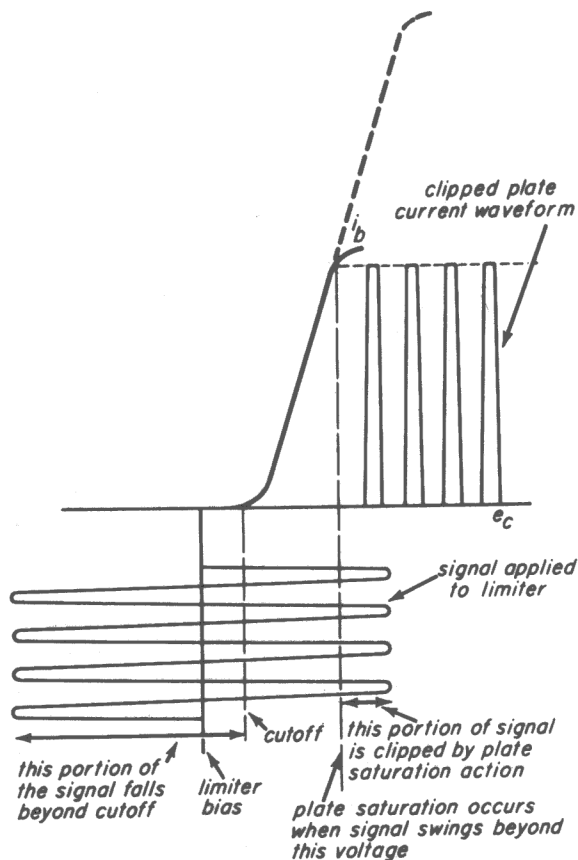


Fig. 37-55

plitude variations that affect the peak of the negative half cycle of the input signal are thus not reproduced in the limiter output circuit, since the tube is cut off when they are taking place.

A sharp cutoff tube is used at the limiter. This type of tube is readily cut off by a relatively small level of input signal.

How the Positive Peaks of the Signal are Clipped. The positive peaks of the incoming signal are clipped by a process known as *plate saturation*. The plate saturation process prevents plate current from rising when the positive-going portion of the input signal voltage has reached a certain level. Even though the input signal rises above this level, the plate current does not; no current or voltage variation takes place in the plate circuit at this time, as a result, and this portion of the input signal remains unreproduced – it is clipped.

In order to attain plate saturation easily, a low value of plate voltage and screen grid voltage must be used. The extended dashed line shown in Fig. 37-55 is the e_c-i_b curve for normal values of plate and screen-grid voltages. The solid line is the e_c-i_b curve for a limiter tube operating with a low value of plate and screen-grid voltages. The lower the B+ voltages, the smaller is the input signal required for the tube to reach saturation.

Limiter Output Signal. As we learned, even though the limiter input signal is clipped, output signal voltage waveform doesn't show this clipping. The waveform of the plate signal is a sine wave, just like the grid signal voltage waveform, except that waveform of the plate signal is of constant amplitude.

The flywheel effect of the tuned circuit at the plate of the limiter is responsible for re-shaping the clipped signal into sine-wave form. Even though the current input to a tuned circuit is not a sine wave, the voltage output of a tuned circuit varies sinusoidally.

Need for Large Input Signal. Satisfactory limiting will take place only when the input signal is large enough to drive the limiter

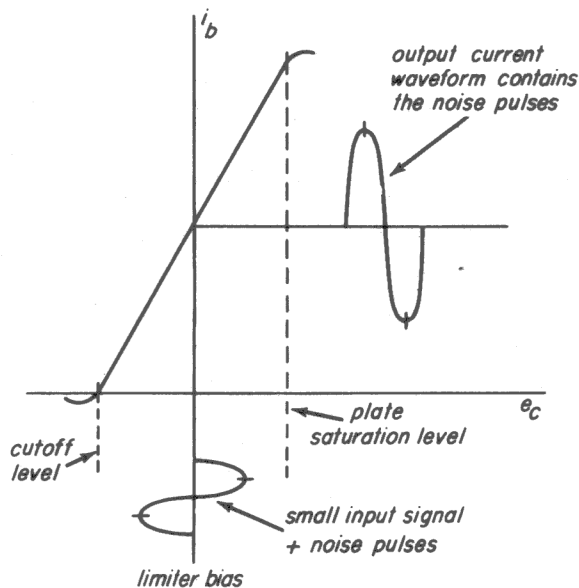


Fig. 37-56

from plate current cutoff to plate saturation (Fig. 37-55). At lower levels of input signal, the limiter functions as an amplifier (Fig. 37-56).

The i-f stages must have a large enough gain, and the antenna a high enough signal pickup, to feed a signal of the necessary amplitude to the input of the limiter. When reception conditions are favorable, that is, when relatively little noise is associated with the transmitted FM signal, a signal with a peak-to-peak voltage of about 2 volts and greater is required at the grid of a limiter circuit for limiting action. When considerable noise is associated with the input signal, this signal will have to be approximately 10 to 20 volts p.p., to produce the large amount of limiting needed to remove its large noise-caused amplitude variations.

The higher the gain of the stages preceding the limiter, the smaller the signal that will be properly limited. With inadequate receiver gain, reception of weak stations is likely to be marred by noise.

Limiter Grid Time-Constant Considerations. The time constant of the limiter grid circuit plays an important part in the operation of the limiter. Actually, there are two time constants: the charging time constant, the time it takes C_1 to charge through the cathode-to-grid resistance of the limiter

tube when grid current flows; and discharge constant, the time it takes for C_1 to discharge through R_1 when grid current has stopped flowing.

The charge constant can be changed only by varying the value of C_1 , since the cathode to grid resistance of the limiter tube remains constant. As a rule, without taking the discharge time constant into account, C_1 is chosen to be as short as possible. With a short charging time constant, the capacitor will reach full charge in a briefer interval than it would with a long constant. It is desirable to have the capacitor reach its full charge potential as soon as possible.

However, the discharge time constant is not so simply determined. It must be long enough so that the limiter circuit will respond to changes in carrier amplitude. On the other hand, the time constant must be short enough so that the bias will change promptly to offset unwanted amplitude variations.

A suitable compromise is reached by means of choosing a time constant that will result in plate saturation of positive peaks of the weakest carrier for which the set is expected to provide good reception. For these conditions, the grid bias can be considered to be cut off or slightly greater.

Once this limit is established, there will be effective limiting of all signals of greater amplitude than the signal just described. If a signal of greater amplitude drives the bias of the limiter more negative the positive peaks of the stronger signal will be sufficiently high to be clipped.

As shown in Fig. 37-57a, if the amplitude of the carrier is increased, amplitude of the noise will fall beyond cutoff and saturation even if the grid-leak bias voltage does not increase quickly in step with the increase in the carrier amplitude, therefore limiter operation will not be impaired. However, suppose the amplitude of the carrier is reduced due to rapid cancellation by the noise called *rapid fading* of the carrier. If the discharge time constant ($R_1 \times C_1$) of

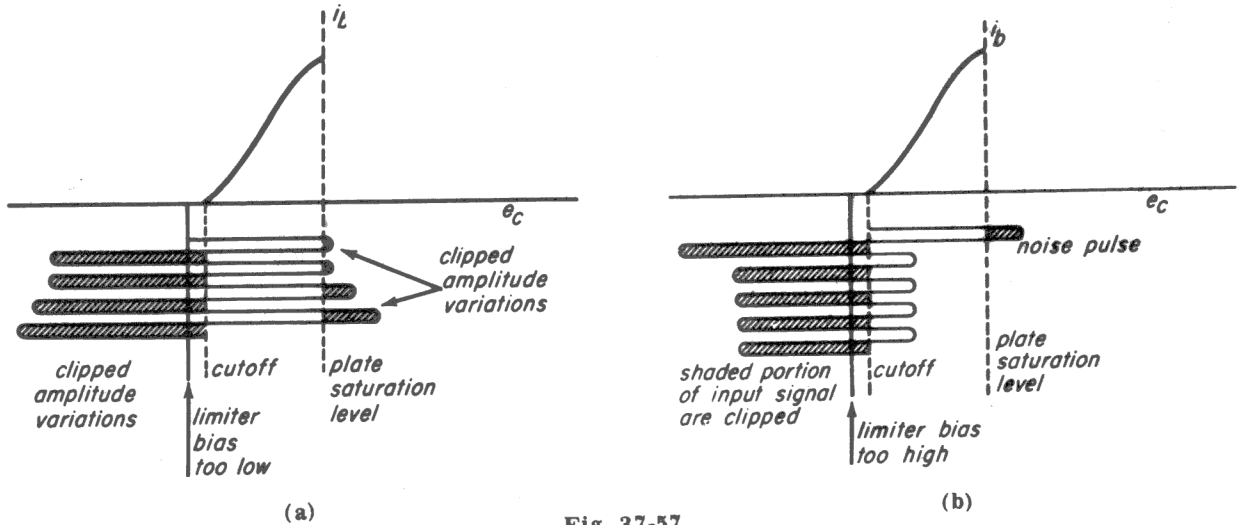


Fig. 37-57

the limiter grid circuit is short enough, the grid-leak voltage will quickly change to a level proportional to the new carrier level and the grid-leak bias will fall.

It is important that the discharge time constant of the limiter be short enough to produce a signal bias voltage that is proportional to the amplitude of the signal and not the amplitude of noise pulses. It must not be *too* short. However, the shorter the discharge time constant, the faster C_1 discharges through R_1 and the smaller the grid-leak voltage becomes. If the time constant is too short (C_1 or R_1 too low in value), the grid-leak voltage will shift with a momentary noise pulse. As a result, the noise pulse will be clipped but the peaks of the carrier will be lower than the noise and will not be clipped. Thus, the entire carrier and a portion of the noise pulse above the carrier will be passed by the limiter (Fig.

37-57b). Circuit operation will be impaired for other reasons as well.

A time constant that is short enough to prevent extra bias voltage from developing due to one type of noise may not be short enough for another type. Many types of noise will be adequately eliminated when the discharge time of C_1 is somewhere between 10 and 20 μsec . For satisfactory rejection of ignition noises, on the other hand, the time constant will have to be much smaller. The time constant actually chosen, thus, must be a compromise value; that is, if a single stage of limiting is used.

To obtain improved limiting, two limiters are employed. They are connected in cascade; that is, one following the other. A typical double limiter circuit is shown in Fig. 37-58. Each has a different time con-

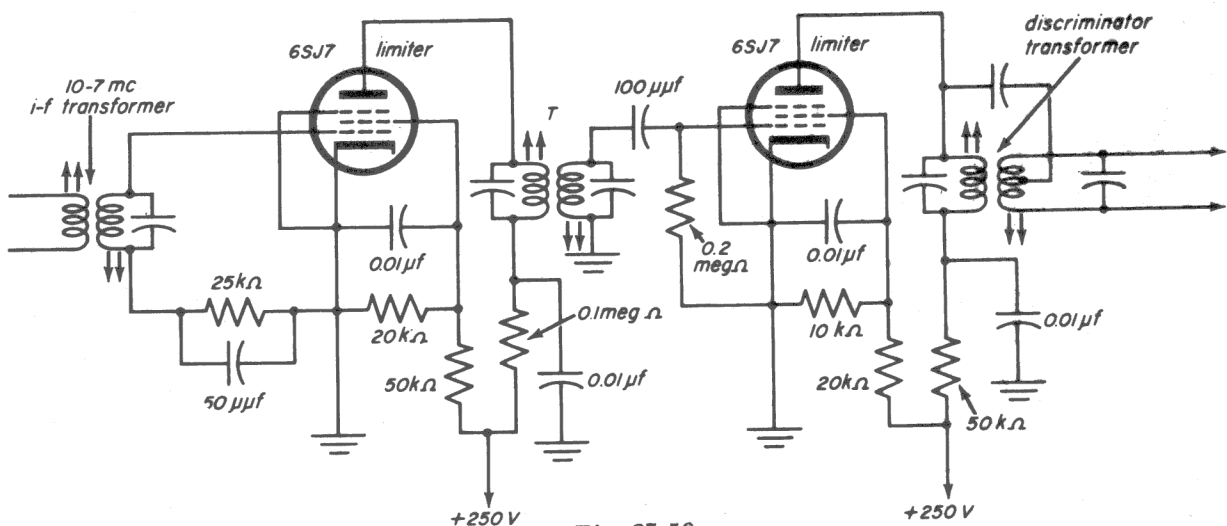


Fig. 37-58

stant in its grid circuit: one short, the other much longer. The first limiter may, for instance, have a discharge time constant of $2.5 \mu\text{sec}$; the time constant of the second may be $10 \mu\text{sec}$ or longer. Better rejection of the different types of noise that may affect the signal is produced through the use of two limiters. A higher i-f signal output from the limiter section is also provided, resulting in louder sound volume (compared with the sound volume that would be heard if only one limiter were present).

Need for Correct Receiver Alignment to Obtain Good Limiter Action. Proper alignment of the tuned circuits in the stages preceding the limiter is essential to obtain adequate noise rejection action from the limiter. If the alignment is slightly improper, that is, if the carrier frequency does not fall at the exact center of the tuned circuit response, the signal frequency furthest removed from the carrier will fall at a point along one slope lower down than it normally would (Fig. 37-59). The difference in amplitude between this signal and the signal that falls at the top of the curve is greater than it would ordinarily be. The result is a larger variation in the amplitude of the incoming i-f signals.

Such an amplitude variation occurring at the grid of an amplifier tube is amplified by the tube, as well as by other tubes following this one, so that it becomes larger yet; the net amplitude variation in the signal at the grid of the limiter may, therefore, be considerably greater than normal. If noise amplitude-modulates this varying input signal, the range of variation in amplitude is increased still further. The net variation at the grid of the limiter that results may be

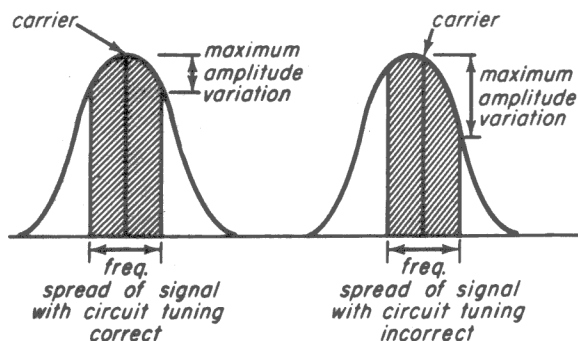


Fig. 37-59

too large to be completely removed by limiter action. Proper receiver alignments and tuning in of the FM signal are required to make the limiter stage operate effectively.

AVC From Limiter. The voltage developed across the limiter grid resistor is proportional to the strength of the incoming signal, and may therefore be used for AVC purposes. When AVC voltage is derived from the limiter, the limiter grid voltage is applied through an isolating resistor (Fig. 37-58) to the grids of the stages controlled by AVC. Use of an AVC system reduces amplitude variations in the incoming signal to some extent.

37-7. I-F AMPLIFIER SECTION

In an FM receiver, i-f amplifiers pass and amplify the desired i-f signals and reject undesired signals whose frequency falls close to the frequency range of the desired signals.

Passing Desired Signals. The incoming i-f signals have a range or spread of 150 kc. The bandpass of the i-f stages, however, must be considerably greater than 150 kc to allow for oscillator drift. If the i-f bandpass was merely 150 kc, any slight drift of the oscillator would cause the i-f signals at one extreme of the bandpass to shift outside the limits within which fairly uniform tuned circuit gain takes place (Fig. 37-60a, b).

Therefore, the i-f bandpass is made 200 kc or more (Fig. 37-60c). As a result, even if there is normal oscillator drift, the shift of the i-f signals will still produce a fairly uniform response (Fig. 37-60d). Making the i-f bandpass wider than 150 kc also compensates for inaccuracies in the tracking of the front end tuned circuits and in the calibration of the tuning dial. Such inaccuracies, like oscillator drift, tend to displace the i-f signals with respect to the bandpass of the i-f tuned circuits.

If the bandpass of the i-f tuned circuits is too narrow to pass the FM signal and all its significant sidebands without undue at-

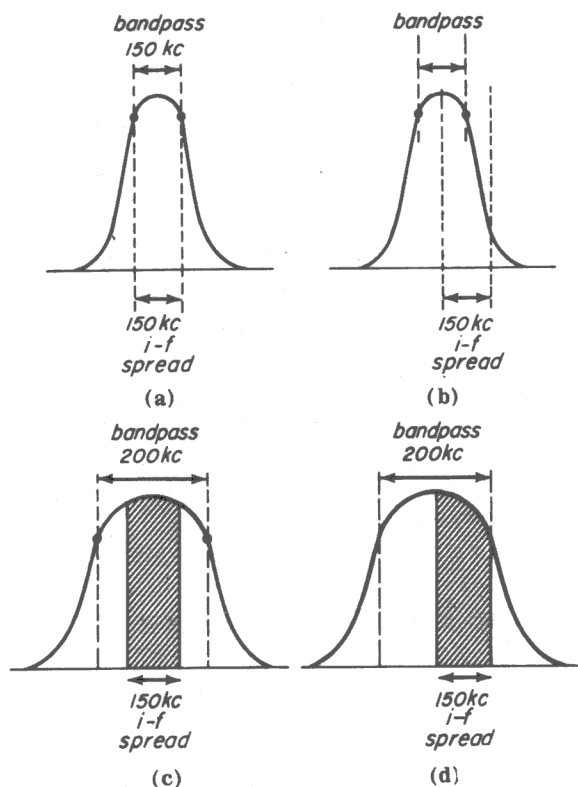


Fig. 37-60

renuation, there will be undesirable effects, chiefly amplitude modulation and phase modulation.

Some amplitude variation of the signal is introduced due to the curvature of the response of a tuned circuit. When the response of a tuned circuit is too narrow, the signals that arrive at the extremes of the response will fall lower down than they should on the response curve (Fig. 37-61a) causing the difference in amplitude between these signals and the signals that come in at the top of the response curve to be greater; the amplitude modulation of the FM signal is therefore greater. However, when the amplitude modulation produced is not too great (Fig. 37-61b) the limiter is capable of removing it.

A much more serious result of an insufficiently wide bandpass is phase modulation. When a signal is coming in whose frequency is the same as the resonant frequency of the tuned circuit, the capacitive and inductive reactances of the tuned circuit cancel, making the circuit resistive only. No shift in the phase of the signal passing through the tuned circuit occurs. But when the signal entering the tuned cir-

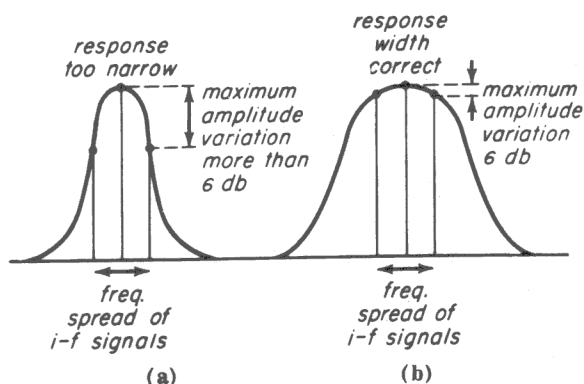


Fig. 37-61

cuit is different from the resonant frequency, the inductive and capacitive reactances do not completely cancel. Some net reactance remains. It will be either capacitive or inductive, depending on whether the frequency of the applied signal is above or below the resonant frequency.

The net reactance reduces the circulating current flowing through the tuned circuit, and lowers the tuned circuit "gain". This results in a smaller amplitude of response at the applied frequency. *The net reactance in the tuned circuit also causes a shift in the phase of the applied signal.*

The greater the separation of the incoming signal from the center frequency of the tuned circuit, the greater the net reactance the signal encounters in the tuned circuit, the smaller the current that circulates through the tuned circuit, the lower the tuned circuit gain, and the lower down on the response curve the signal falls. If the reactance encountered by the incoming signal is greater, the phase shift the signal undergoes is larger. Therefore, signals that fall at a low point on the response also undergo a certain amount of phase shift.

The amount of reactance a signal encounters in a tuned circuit is indicated by its relative amplitude on the tuned circuit response compared with the peak response amplitude. If the amplitude difference is small, the signal is meeting little more reactance than at the center frequency and is undergoing little phase shift. If the amplitude difference is large, the signal is meeting considerably more reactance than the center frequency and is being subjected to a relatively large phase shift.

Phase shift is equivalent to a small frequency deviation. Such a frequency deviation is undesired. Unfortunately, it cannot be eliminated. The undesired frequency deviations produce undesired voltage outputs from the FM detector. If these voltages are appreciable, noticeable distortion results.

The distortion is most evident at high-amplitude audio signals. I-F signals farthest separated in frequency from the carrier represent these high-amplitude audio signals. These i-f signals come in at points along the tuned circuit response farthest removed from the center point, or at levels on the response lower than those of other signals. If these levels are much lower than the center point, signals that come in at these levels will encounter relative large amounts of net reactance in the tuned circuits, suffer large amounts of phase shift, produce relatively large frequency deviations and large voltage outputs from the FM detector, and thus introduce audible distortion.

When the bandpass of a tuned circuit is adequate, the difference between the center of the response, and the points on the response at which the signal frequencies furthest removed from the center come in, is approximately 6 db (Fig. 37-61b). The maximum phase shift introduced as a result of such a range in response amplitude does not introduce noticeable distortion. When the bandpass of the tuned circuit is too narrow, however, the difference in amplitude between the response levels at which the outermost (maximum deviation) signals fall and the center of the response is more than 6 db. As a result, there is too much phase shift at the outermost signal frequencies.

Rejecting Undesired Signals. The i-f signals produced by channels adjacent to the one that is being received may be strong enough to get through the tuned r-f circuits of the receiver. The response of an i-f tuned circuit must slope down at its sides sharply enough to produce a negligible output at these unwanted intermediate frequencies.

It is desirable to make the intermediate frequency low, since a low value of i-f

makes possible a relatively high gain per stage. But if the i-f is made too low, image interference becomes possible.

Both the desired and the undesired i-f signals will be accepted by the i-f tuned circuits. The undesired signal is called the *image signal*. The interference it produces is known as *image interference*.

To prevent image interference from signals in the FM band, the intermediate frequency must be such that the image frequency falls outside the FM band at any setting of the tuning dial and is outside the bandpass of the tuned circuits in the r-f amplifier and mixer. This condition will be satisfied if the i-f is made slightly more than one-half the width of the FM band. The FM band is 98 to 108 mc, or 20 mc wide. One-half the FM band is 10 mc. The i-f commonly used in FM receivers is 10.7 mc, which is slightly more than 10 mc. The r-f image frequency is always separated from the desired r-f signal by twice the i-f. With an i.f. of 10.7 mc, twice the i.f. will be 21.4 mc. No matter what channel is tuned in, the image will always be 21.4 mc away, which will put it outside the FM band at any setting.

Image interference from signals outside the FM band remains possible. The selectivity of the r-f tuned circuits is relied on to prevent such interference.

Tuning Methods. The i-f circuits may be tuned by means of a variable capacitance or a variable inductance. Variable inductance tuning is preferable to variable capacitance tuning, and is used more often. Let's consider the advantage of varying the inductance of the coil in the tuned circuit, rather than the capacitance of the capacitor.

Engineers have proved that the gain of a stage depends, in part, on the L/C ratio of the tuned circuit components—i.e., the ratio of the coil's inductance to the capacitor's capacitance. The higher this ratio is, the greater the gain of the tuned circuit, and the gain of the stage. The L/C ratio may be made larger by increasing L or decreasing C.

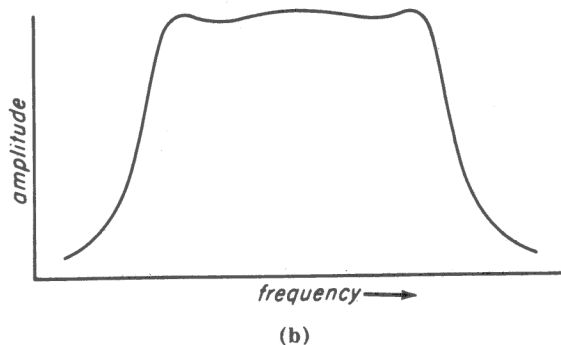
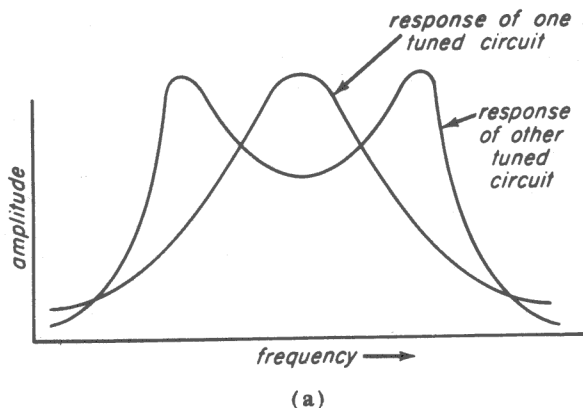


Fig. 37-62

When variable capacitance tuning is used, C cannot be made very small, since a relatively large capacitance must be used, to permit the proper amount of tuning variation. When a variable inductance is used, on the other hand, C can be reduced to a small value in order to provide a suitable large L/C ratio. Making the coil inductance large enough to provide an adequate tuning variation introduces no big problem, since the L/C ratio is increased, not reduced, by using a large value of inductance.

Variation of the inductance of a coil is commonly produced through the use of a method known as *permeability tuning*. A threaded slug made of powdered iron, held together by a plastic binding material, is inserted into the coil. Moving the slug all the way into the coil increases the inductance to maximum; drawing it as far out as it will go reduces the inductance to minimum. At in-between settings of the slug, values of inductance intermediate between minimum and maximum are obtained.

Coupling Methods. The bandwidth of an i-f stage is to a large extent determined by the method of coupling between i-f stages. Several types of coupling have been employed in FM receivers. One uses double-tuned transformer coupling similar to that present in the i-f stages of an AM receiver. To broaden adequately the bandpass of the tuned circuits, a resistor of suitable size is often connected in parallel with the capacitor and coil. The resistor lowers the Q of the tuned circuit and thus increases the bandpass.

In some instances, the tuned circuits are resonated at different frequencies. The combined response of the circuits overlap, making the over-all response suitably broad and flat (Fig. 37-62).

I-F Transformer System for FM-AM Operation. Many receivers are designed to receive both FM and AM stations. For reasons of economy, receivers of this type use the same i-f tubes to pass FM i-f and AM i-f signals. A typical i-f transformer circuit designed for such operation is shown schematically in Fig. 37-63.

The upper tuned circuits are used to pass FM i-f signals; the lower ones pass AM i-f signals. The AM i-f primary and FM i-f primary windings are connected in series; the same is true of the secondary windings.

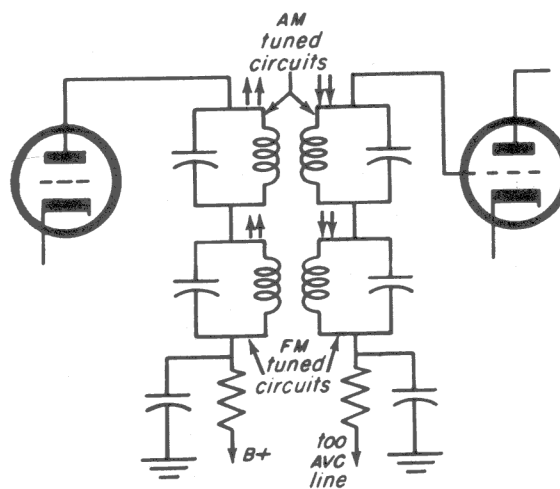


Fig. 37-63

The two sets of windings are sometimes connected in the same can; more commonly, separate cans are used for each set of windings. The advantage of enclosing both FM and AM windings in the same can is economy. The disadvantage is that the FM and AM windings interact to a greater extent than if they were contained in separate cans, making receiver alignment more difficult.

On FM operation, the capacitor of each 455-kc AM tuned circuit provides a short-circuit path for the 10.7 mc signal, shorting out the AM tuned circuit and preventing an undesired FM i-f signal voltage from developing across it. On AM operation, the inductive reactance of the coil in each FM

tuned circuit is effectively a short-circuit for the AM i-f signal, preventing the AM i-f signal from developing an appreciable voltage across the FM tuned circuit.

The i-f tubes used have a high transconductance, because high gain in the i-f stages is needed for satisfactory FM reception. A comparable degree of gain is *not* needed for the incoming AM i-f signals. Since the same tubes are used to amplify both the FM and AM i-f signals, steps must be taken to keep the AM i-f signal output from being excessive. The necessary reduction in gain is achieved by using a very low L/C ratio in the 455 kc i-f tuned circuits.

